Proceedings of the I·R·E

JOURNAL of the Theory, Practice, and Applications of Electronics and Electrical Communication

adio Communication • Sound Broadcasting • Television • Marine and Aerial Guidance • ıbes • Radio-Frequency Measurements • Engineering Education • Electron Optics • ound and Picture Electrical Recording and Reproduction • ›wer and Manufacturing Applications of Radio-and-Electronic Technique • dustrial Electronic Control and Processes • Medical Electrical Research and Applications •

OMBINED OPERATIONS

(Shoulder Flash)

/orn by all personnel serv- g permanently under Com- ined Operations Command. ıey include not only the itish "Commandos" and en of the Royal Navy and e R.A.F., but also U. S. angers and personnel of ther Allied forces.

FEBRUARY, 1945

VOLUME 33 NUMBER 2

Electronics in Industry

U-H-F Multiplex Radiotelephony

Studio-to-Transmitter Antenna

Reflex Oscillators

Theory of Transmission Lines

NATIONAL SAVINGS AUTUMN DRIVE

WE'VE GOT TO KEEP ON SAVING!

The Institute of Radio Engineers

Proceedings *of the* I·R·E

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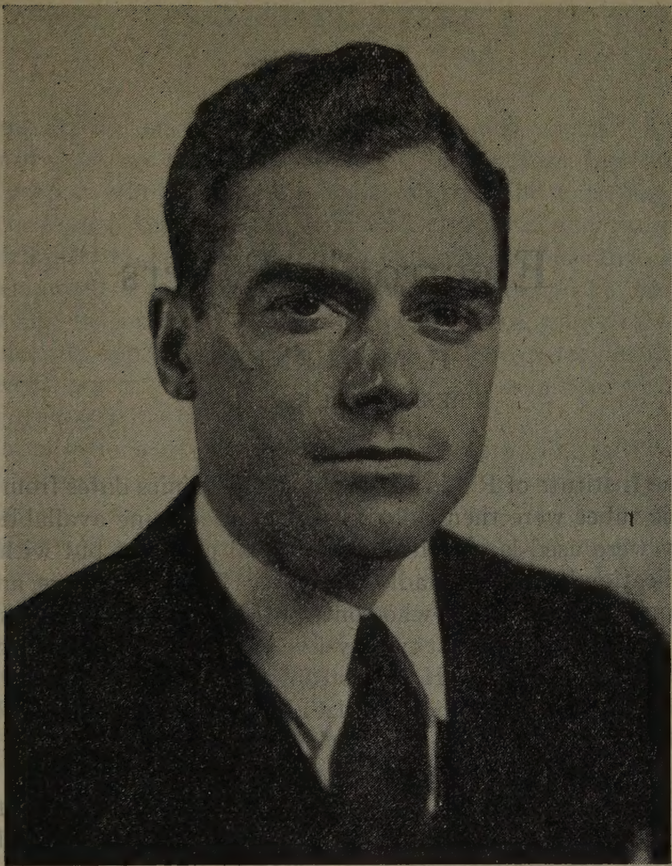


Browder J. Thompson

1903–1944

Among the most unfortunate and sorely felt effects of the war upon The Institute of Radio Engineers are the inevitable casualties amongst its membership. Browder J. Thompson, killed in action during a special mission for the Secretary of War, is one of these. Mr. Thompson lost his life on the night of July 4–5, 1944, during a flight in an Army plane in the Mediterranean Theater, while serving as expert consultant in the Office of the Secretary of War and performing a mission described as "of direct and vital importance to the war."

Born on August 14, 1903, in Roanoke, Louisiana, Mr. Thompson received the Bachelor of Science degree in electrical engineering from the University of Washington, Seattle, in 1925. In 1926 he entered the research laboratory of the General Electric Company, where he was engaged in the design and development of vacuum tubes for radio and industrial use. In 1931 he was transferred to the RCA Radiotron Company, in Harrison, New Jersey, in charge of the electrical research section of the research and development laboratory. As a result of his outstanding work with that organization in the development of vacuum tubes of small physical dimensions for ultra-high-frequency uses, Mr. Thompson was awarded the Morris Liebmann Memorial Prize for 1936, with the citation "for his contribution to the vacuum-tube art in the field of very-high frequencies."

From 1942 until he accepted his Government assignment, in December, 1943, Mr. Thompson was associate research director of RCA Laboratories, Princeton, New Jersey. He was recognized as one of America's foremost radio research engineers, including amongst his contributions such important work as guiding the development of the famous "acorn" tube used in ultra-high frequencies, and advances in screen-grid tubes and power pentodes that became mainstays in broadcast reception.

Joining The Institute of Radio Engineers as an Associate member in 1929, he was transferred to Member grade in 1932, and became a Fellow of the Institute in 1938. His association with this organization was a very active one; he served on numerous committees, participated in Standards work, and was a member of the Board of Editors and of the Board of Directors from 1937 until his death. Mr. Thompson was also a member of the American Physical Society, Tau Beta Pi, and Sigma Xi.

As one who had a consistently outstanding record of service to his field and an upright and co-operative personality, Mr. Thompson will be sincerely missed by his friends and associates.

Electronic Papers

H. M. TURNER
PRESIDENT, I.R.E., 1944

The interest of The Institute of Radio Engineers in electronics dates from the early days. The audions, as electronic tubes were then called, were just becoming available when the Institute was organized. They were used largely for experimental purposes but within a few years they found extensive application in wire and radio communication as detectors, amplifiers, oscillators, and modulators. Many of the engineers who contributed to this development have received recognition by the Institute during the past twenty-five years. Among them may be mentioned the following: Medal of Honor: Major E. H. Armstrong, 1917; Dr. Lee de Forest, 1922; Dr. J. A. Fleming, 1933. Morris Liebmann Memorial Prize: R. A. Heising, 1921; J. R. Carson, 1924; Dr. A. W. Hull, 1930; Stuart Ballantine, 1931; Heinrich Barkhausen, 1933; Dr. V. K. Zworykin, 1934; Dr. F. B. Llewellyn, 1935; B. J. Thompson, 1936; W. H. Doherty, 1937; P. T. Farnsworth, 1941.

The increase in the number of electronic papers published in the PROCEEDINGS during the past year has met with the approval of our members. As the techniques developed in connection with the communication art and research continue to find applications in many diverse fields, there has arisen a need for a broader coverage of the subject than has been the practice in the past. Engineers having electronic experience are called upon to apply their knowledge outside the communication field, and in the interest of efficiency they must be fully informed as to what is being done by others. To meet this need the Board of Directors has recently approved a further increase in the emphasis on electronics and has greatly extended the subject material acceptable for the PROCEEDINGS. Some indication of this policy is given by the topics listed on the outside cover of this issue, and more in detail in the editorial by Dr. Goldsmith in the December, 1944, issue in which prospective authors of electronic papers are invited to present them for publication. I urge the co-operation of authors in order that the PROCEEDINGS may cover all phases of electronics adequately and authoritatively.

The Institute Looks to the Future

WILLIAM L. EVERITT
PRESIDENT, I.R.E., 1945

"Electronic devices are not small inherently, they are only young." This intriguing remark was made a number of years ago by Dr. A. W. Hull, a Fellow of the Institute and pioneer worker in electronics. The prophesy implied therein has come true, for many electronic devices are no longer either small or young. But the remark gives a clue as to why radio and communication engineers have contributed to, and are interested in, such a broad range in scientific and technical research, development, and application. Because devices can be used in radio when they are small, they can also be used when they are young. Hence, many new scientific discoveries find their first engineering applications in the radio field. The history of the application of electricity to human needs, starting with the Morse telegraph, is replete with examples where the communication engineer was the pioneer.

But engineering applications grow both in size and breadth. The radio engineer, and his professional society, The Institute of Radio Engineers, have a continued interest in the fields of activity which grow out of his work. The Institute welcomes to its membership the co-workers who are engaged in further research and development in all these expanded fields. In the Institute, they will find the largest group with an interest common to their own, and the publication medium which will give their work the most appreciative audience.

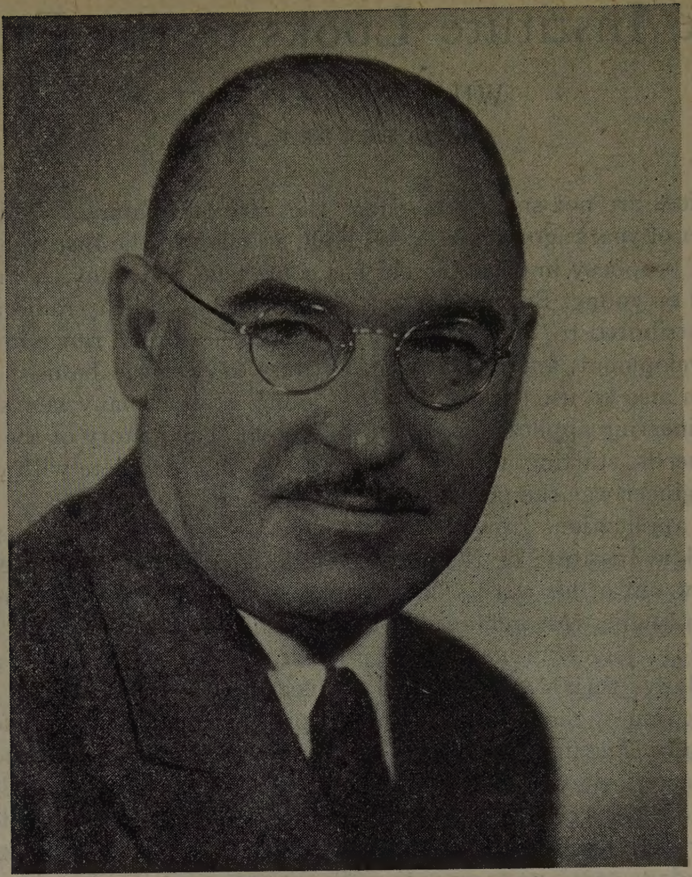
No better illustration of the extensive use of a principle first developed for radio use can be found than electronics. The first application of an electronic device, as it is now defined, was the Fleming-valve detector, whose usefulness was greatly enlarged by the addition of the grid by de Forest. The applications of electronic devices have now expanded so widely that a mere list of uses would cover many pages. The Institute of Radio Engineers has also grown with the electronic art and intends to continue to grow to serve the entire family of engineers who are working in this field. It has shown, and will show, this growth and broadening by expansion of its publication policy, by development of its technical committee functions, by coverage in the technical meetings of the Sections and the Institute at large, and by such other means as the membership may desire and will support.

The programs for securing permanent headquarters and for enlargement of the permanent staff have all been considered by the Board of Directors as a means for serving the broad fields in which the Institute is the natural leader. This Board is composed of your representatives. In performing its functions it needs the advice and suggestions of the members. This advice will be most useful if it is specific in its proposals, so that concrete action may be obtained as rapidly as possible. Such advice is definitely solicited. Plans which are adopted will also require *work* on the part of the members to put them into practice. Only by concerted, co-operative effort can any professional organization progress and meet the needs of a changing world.

The Institute has adopted and confirmed a membership structure whereby those in professional engineering practice can hold voting membership. It is important that those qualified should avail themselves of this privilege by transferring to the appropriate grades. Each Section is being asked to form a local committee to aid in securing sponsors and otherwise assist those who should transfer to the higher grades. It is important that this work should go forward rapidly and energetically, so that all the professional members can have a part in guiding the policies of the Institute.

It is also important that the Sections should seek out and secure members from workers in the electronic and allied fields, and that they should arrange technical programs which will serve such members. This is not a new policy. Many Sections have done it for years, but we are reemphasizing our interest, and taking positive action to implement it throughout all the activities of the Institute.

The coming years present a great opportunity. We must and will meet it.



Raymond F. Guy

Raymond F. Guy became interested in radio as an amateur in 1911. From 1916 to 1917, he served as Marconi Wireless Telegraph Company ship's radio officer for one and a half years, when he resigned to enlist in the Army and served in France.

Upon discharge he entered Pratt Institute, graduating in electrical engineering in 1921. A vacation was spent with the Independent Wireless Telegraph Company, and later he entered the Shipowners Radio Service.

In 1921, Mr. Guy began twenty-four years as a broadcast engineer, at Westinghouse Station WJZ, Newark, when it opened as the world's second broadcast station. Commercial broadcasting did not exist, the audience consisted of only a few amateurs, and practically all apparatus and techniques had to be conceived and developed.

In 1924, he was transferred to the RCA research laboratory and later was made head of the broadcast engineering section which engineered the RCA-NBC stations, did consulting work for clients, and directed development of all broadcast transmitting apparatus used or sold by RCA. During this period, he collaborated in the first trans-Atlantic rebroadcast and the development of the service, improved and extended network broadcasting, and collaborated in the design and construction of many broadcast plants in the United States, Europe, and the Orient.

In 1929, he transferred to the National Broadcasting Company to become NBC's radio-facilities engineer. During the last fifteen years he has directed the planning, design, and construction of all of NBC's radio facilities for standard broadcasting, frequency modulation, television, and short-wave broadcasting throughout the United States, and his duties have included every phase of engineering and economics associated with such facilities.

Mr. Guy has taken an active part in establishing the position of the United States in short wave broadcasting since 1925, when he participated in the design and construction of a 40-kilowatt RCA plant at Bound Brook, New Jersey. Under his supervision NBC has built nine stations, one of them of 200 kilowatts power.

Mr. Guy became an Associate of the Institute of Radio Engineers in 1925, a Member in 1931, and a Fellow in 1939. In 1943, he was elected to the Board of Directors for 1944–1946. He is a Fellow of the Radio Club of America, a charter member of the Twenty Year Club, a member of the New Jersey Society of Professional Engineers, a life member of the Veteran Wireless Operators Association, and is admitted to practice as a Professional Engineer in New York and New Jersey.

He serves on four panels of the Radio Technical Planning Board, is secretary of a Committee of Panel 4, and chairman of the International Broadcasting Committee of Panel 8. He has served on the I.R.E. Broadcast Committee, the Technical Committees on Transmitters and Antennas (as chairman), as a member of the Standards Committee, two New York Section Committees, is 1944 chairman of the Transmitter Committee, and is a member of the Committees on Frequency Modulation and the Annual Review. He also served on a Technical Committee of the American Standards Association.

Electronics in Industry*

W. C. WHITE†, FELLOW, I.R.E.

Summary—The science of electronics is considered one of the hopeful influences to help in maintaining a high rate of industrial activity in the postwar era.

To do this requires that new electronic applications for industry be developed rapidly. A review of such applications in the past indicates a period of transition from initial laboratory work to extended commercial use that is too slow to be helpful in any immediate postwar period. The reasons for this are considered.

Under the impetus of war, this period of time required for development has been cut drastically in the case of electronic devices for military use. The reasons for this speeding up are also considered.

In conclusion, there is reason to believe that some of the wartime speeding-up influences will continue on into the postwar era. These, in combination with previous experience and background, give encouragement to the belief that electronics in industry will be of real help in creating new postwar business activity.

NO ONE here needs to be told that the war has expanded the radio and electronics business to an extent matched by few, if any, other of our industries. Also, no one needs to be reminded that there is certain to be a tapering off of the now huge volume of business in electron tubes and electronic devices in the initial postwar era. To complicate the problem further, the war requirements brought many new concerns into this field as well as causing the marked expansion of many former ones already established.

There is much hope that all of the new ideas and developments that have resulted from the war effort can be put to use to produce new devices to minimize the postwar recession in business volume. The big question and likewise the big problem involved is how quickly some of these new ideas and developments can be brought into commercial use.

As my assigned subject is industrial electronics, I shall talk only on this phase of the subject, and shall not include the extension of broadcasting that will result from the development of frequency-modulation stations and the new science of television.

The Past

In the industrial application of electronics, I have always been impressed, as well as distressed, by the amount of time taken for the transition between a good idea or a laboratory development and active commercial utilization. It is natural to ask the question: "Why has it taken so long successfully to introduce new electronic devices and methods?" Let us take, for example, three electronic devices which are really not new, but which have come into extended use only quite recently:

1. *High-Frequency Induction Heating With Electronic Tubes.*

I remember building one of these equipments of about ½ kilowatt output as early as 1919 and being impressed with its capabilities. I fondly believed that, within four or five years, the use of this principle in industry would be quite widespread. However, it was nearly 20 years before this business amounted to very much. If one looks at the many units being installed today, one is impressed by the fact that they could just as well have been built ten years ago as regards cost, design engineering knowledge, and the availability of the components.

2. *The Operation of Direct-Current Motors from an Alternating-Current Supply System through the Medium of Thyratron Control.*

Here again, the tubes, components and general circuit arrangements were well-known ten years ago, but it has been only during the last few years that equipment of this type has been used to any great extent.

3. *Rectifiers to Supply 250 to 600 Volts Direct-Current for 25 to 500 Kilowatts Utilizing Sealed-Off Ignitron Tubes.*

It has taken a number of years to establish this type of rectifier in the electrical field. No new major technical elements have entered the picture since the advent of the ignitron in 1933.

If one looks for certain basic reasons for the long period of development in cases such as these, a number of factors are apparent. The most important of these seem to be:

(1) The cynicism and stagnation of initiative that accompany a business depression. With an atmosphere dominated by this attitude, it is very difficult to launch new ideas quickly.

(2) There is another psychological factor: the mental resistance to change, which is a most powerful influence. Probably the best example of this in the electronics business was the difference in attitude five or ten years ago, that existed in England and the continent as compared with this country, toward sealed-off tubes and medium-power rectifier units. This particular branch of electronics was much more advanced abroad than in this country. As we were naturally interested in promoting this business, we went to considerable pains to find out why this country lagged so far behind in this particular application. Although there were a few technical factors, they were relatively minor, and the real cause seemed to be that the electrical engineers abroad were rectifier-minded and used rectifiers rather than motor generators for this particular class of application. In other words, abroad it had passed through the various developmental stages and had become an established business. Still another way of expressing it is to say that

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many engineers are habit-bound. The manufacturers of industrial electronic equipment have always found that their most difficult competition came, not from other manufacturers of similar equipment, but from the old way of doing things.

(3) Somewhat related to this last item is what has aptly been called the "vicious circle in new developments." It is very difficult to design and build a thoroughly satisfactory device without considerable field and operating experience, but it is also difficult to get such operating experience in the hands of a user unless the device is thoroughly satisfactory. Sometimes new designs are plagued by certain little detail complications or troubles that exasperate the pioneer user and color his whole attitude toward the device, overshadowing many of its advantages and good features. This often makes prospective users loath to try or keep a device.

(4) A survey we made some time ago indicated that first cost and maintenance expense were important factors which deterred potential users of electronic equipment from adopting such devices. Unlike radio equipment, which absolutely requires an electron tube in order to function at all, industrial electronic equipment must usually compete with older and often highly-developed methods. The fact that these older methods are highly developed usually means that they have received a great deal of attention as regards cost reduction.

(5) The second important factor that our survey indicated was ignorance of what the device would do and how it operated. Until recently, electronic devices were thought of as something different and often mysterious. Service men handling industrial electronic devices were at times called in to repair them, and found that the only defect was something simple and obvious, such as a blown fuse. The real trouble, therefore, was that the users feared the unknown and would not apply the elementary checks and tests that are applied to any other piece of electrical equipment that has stopped functioning.

(6) A trait which industrial electronics inherited from radio, and which in the past proved a considerable barrier to its growth, was the item of relatively short tube life. Only a few years ago, a 1000-hour life for the radio transmitting tubes in a broadcast station was generally accepted. In industry, however, electronic devices must usually compete with other methods. Such a short tube life, which should really be expressed as "high tube-replacement cost," often proved an insurmountable difficulty. Analysis and experience have shown that a 5000 to 25,000-hour tube life is usually essential to enable electronic devices to be competitive with older methods in the industrial field. In general, the larger and more expensive the tube, the longer its life must be for industrial acceptance.

(7) A factor that has presented difficulties to the designer of industrial electronic equipment is the variation in characteristics that exists between individual tubes of the same type. Although the tube engineer has done much in this regard, the fact remains that in such items as transformers, capacitors, resistors, and motors the variation from unit to unit is very much smaller than in the case of tubes. Also tube performance and life vary more with line voltage than for most other basic electrical circuit components. This, too, has imposed special limitations on the designer of electronic devices for industry.

The Present

Let us leave the past and its growing pains and look at the present. Industrial electronic equipments are playing an indispensable role today in America's record-breaking war production. A comparative newcomer to industrial plants, since the major developments in industrial electronics have taken place since World War I, this industrial tool has been directly responsible for saving millions of man and machine-hours, and millions of pounds of critical materials since Pearl Harbor.

No single kind of industrial electronic equipment has contributed more to the building of war machines than resistance welding control. It is being used to fabricate aluminum and many new and special metal alloys which have come into common use. The high quality of welds required to stand the abuses of wartime operations, particularly aircraft, is obtainable only with the precision timing and heat control offered by electronic control.

This war application of industrial electronics and such others as the tank mercury-arc rectifier, automatic controls, X-ray examination of metals, and induction heating will be covered in detail in later sessions of this conference.

As a result of the war, many new military electronic devices have progressed from the idea stage to actual use in a remarkably short period of time, sometimes only a matter of months. In the light of the retarding factors mentioned, it is interesting to study this accelerated pace and try to arrive at the reasons back of it. Many of these are clearly apparent:

1. In the place of the deadening mental attitude that accompanies depression, there is, instead, the atmosphere of energy and a determination to get results and get them quickly.

2. In place of the normal feeling of resistance to change, there is the certain knowledge that, unless we get ahead and keep ahead of our enemies, we are defeated. Thus, necessity for change and improvement becomes a very part of our existence.

3. The so-called vicious circle of trial and error does not exist or is taken in stride as a necessary cycle.

4. Cost and expense are minor factors if the equipment even promises to help win the war. As a matter of fact, very often several alternative methods are developed simultaneously with the knowledge that one or more of them may be a complete failure and be abandoned at an early date.

5. One can only guess as to how many thousands or perhaps tens of thousands of men and women during the period of the war have been given intensive training in the operation, care and maintenance of a multitude of

electronic devices. Much good publicity has also helped to minimize the ignorance factor.

It is obvious that a few of these factors are not just temporary but will carry over into the postwar period and, therefore, represent net gains.

The Future

Now, let us consider the future. In view of the progress that has been made, it is only natural to consider next to what extent, for the immediate postwar era, we can use what we have learned and take advantage of changed conditions in order to shorten the elapsed time between the laboratory and extended commercial use. It would appear that the following factors are important in this respect.

1. Electronics, to the industrial engineer, is another tool to add to his kit for use in conjunction with other well-developed methods. Electronics will not relegate to the background such devices as the induction motor, transformer, amplidyne, and highly-developed instruments. It is clear that, in the final analysis, electronic devices will take their place beside other electrical devices in industry, supplementing and adding new accomplishments rather than superseding many methods now well established.

The successful use of electronic devices in industry is based upon their giving better results than other methods, or because the engineering problem involved cannot be solved in any other way. Practical utility, therefore, is a prime requisite.

Where glamour has been responsible for putting to work some kinds of electronic devices, the chances are almost 100 per cent that these are misapplications rather than progressive steps. There is some reason to believe, for instance, that in a few cases high-frequency heating is today being employed in applications where former types of electrical heating would give satisfactory results and would do it at lower first cost, higher efficiency, and lower operating expense.

2. Another factor that must be kept in mind is that, in most industrial applications, an electronic device is only one part of a larger piece of apparatus. In fact, in the majority of cases the electronic part of the apparatus comprises only a fraction of the cost of the whole. However, in such cases, the electronic part of itself may not only be of very limited use, but the rest of the equipment without its use may also be of little value. With the two parts working together in the proper way, they make a contribution to industrial progress. This leads to the conclusion that electronic devices in industry should not be of the nature of gadgets attached to some piece of equipment, but rather must be engineered as a closely-knit part of the whole. This engineering must be done by men who are familiar with all of the special conditions involved in the particular industry in which the apparatus is to be used, and this factor cannot be supplied by a radio engineer or someone simply familiar with the electronic principles involved.

3. The place where the first trial application is to be made should be one where a high first cost and a high maintenance cost can be justified. Of course, there are entirely too few of these to constitute a real field of business, but usually one can be found that is unusually favorable.

4. Just as important as a trial installation that is economically favorable is one that is in the hands of what might be termed "a friendly customer." The user of a new and untried device in practically every case needs patience, a pioneering viewpoint, an unusual spirit of co-operation, and abiding faith, in order to come through successfully.

5. The developer of the device also must have a great deal of faith, as well as the ability to keep at the development regardless of difficulties, if it is to be really successful.

6. Not only must the tubes have built into them, by design, a long, useful life, but they should be operated conservatively to realize this life. Likewise, users of industrial electronic equipment must be convinced that a few dollars additional first cost in the equipment to permit conservative operation of the components will be repaid many times in lower tube maintenance cost as well as greater freedom from service interruptions.

7. In contrast to radio transmitters and public utility apparatus, there is a minimum of operating personnel whose sole duty is to watch the results, make adjustments from time to time, and see that every item is frequently inspected. To a much greater extent, continuity of operation must be designed into industrial electronic equipment.

All of this is just another way of saying "nothing succeeds like success." Once a new device has shown satisfactory and worth-while service in the hands of a user for a six-months to two-year period, the battle is usually won.

Competition is one of the most powerul of business influences. Let some one industrial concern have a device that lowers cost or produces a better quality of product, and almost over night the other firms in this field will, in the great majority of cases, quickly adopt the idea and be in the market for like equipment.

In closing, I want to remind you that these problems I have discussed are not such that they cast any doubts on the future of electronics. That future is assured. The point is that we want to bring about that coming expansion in the characteristic American way which means: just as rapidly as possible.

Cape Charles-Norfolk Ultra-Short-Wave Multiplex System*

N. F. SCHLAACK†, SENIOR MEMBER, I.R.E., AND A. C. DICKIESON†, SENIOR MEMBER, I.R.E.

***Summary*—This paper describes the general features of a radio multiplex system which has been installed between Cape Charles and Norfolk, Virginia. The radio-frequency equipment operates in the vicinity of 160 megacycles. The system employs the 12 telephone channels of the type K cable carrier system which are in the frequency range 12 to 60 kilocycles.**

General

THE economies afforded by multiplex operation in telephone transmission by wires have always made the attainment of such operation an attractive goal to be achieved in radio transmission. In this type of service, even where directive antennas are used, the radio-frequency power involved is relatively large and is costly to produce. The efficiency, therefore, is of great importance and it becomes impracticable to reduce the intermodulation by operating the vacuum tubes of the output stage of a transmitter at low signal amplitudes. Schemes for spread-band operation have been used to avoid the most serious intermodulation products at the cost of greater bandwidth.[1-3] However, the invention of negative feedback,[4] which can be applied so as to reduce the intermodulation to a satisfactory value at signal amplitudes approximating the rated output of the tube in the final stage of a transmitter, has given the radio engineer a new tool with which to attack the multiplex problem. In addition, the accumulation of further knowledge regarding the propagation of ultra-short waves and the development of more suitable tubes have facilitated the application of multiplex to certain radio transmission problems. This paper describes the general features of an ultra-short-wave multiplex system which was installed in 1941 between Cape Charles and Norfolk, Virginia, as indicated in Fig. 1. It is used by the Chesapeake and Potomac Telephone Company to provide public telephone service over this route.

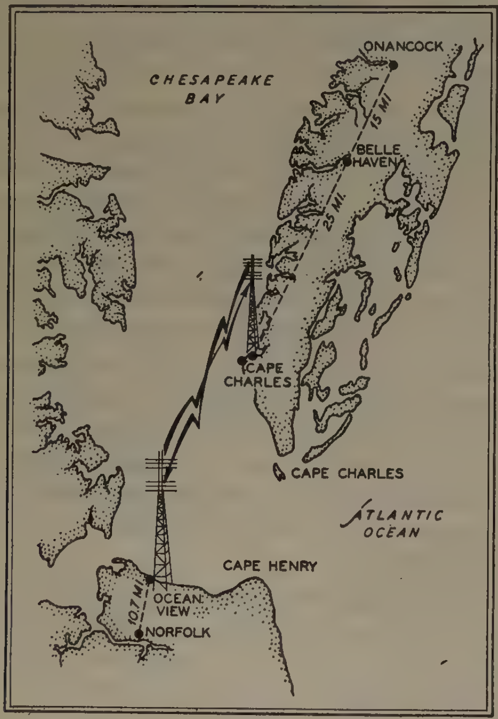

Fig. 1

The path across the mouth of Chesapeake Bay is a particularly suitable one for the application of multiplex radio. Telephone traffic to and from Cape Charles and vicinity has heretofore been handled by wire circuits routed some 450 miles around the bay by way of Baltimore and Washington. By locating radio equipment at Cape Charles and at East Ocean View, Virginia, this circuitous land route has been replaced by a circuit made up of a radio link of 26 miles and only 12.0 miles of wire.

In order that the radio equipment might fit into the telephone system with a minimum of special engineering from a wire transmission standpoint, the radio transmitters and receivers were designed to accept and deliver twelve channels of the type K carrier system[5] which lie in the range from 12 to 60 kilocycles. From a consideration of the various factors involved, it was decided that the radio link should be engineered on the basis that it should be at least as good in performance as 1000 miles of the type K system. This imposes on both the transmitter and receiver unusual requirements in respect to noise and distortion or intermodulation.

At the southern end of the circuit the carrier equipment is installed in the toll office at Norfolk so that the transmission to East Ocean View, a distance of about 11 miles, is accomplished at the carrier frequencies of 12 to 60 kilocycles. At Cape Charles the carrier equipment is located at the radio station so that transmission to the Cape Charles Central Office, a distance of 1.3

* Decimal classification: R460. Original manuscript received by the Institute, August 7, 1944. Presented, Winter Technical Meeting, New York, N. Y., January 27, 1945.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] A. H. Reeves, "The single sideband system applied to short-wave telephone links," *Jour. I.E.E.* (London), vol. 73, pp. 245–279; September, 1933.

[2] E. H. Ullrich, "Ultra-short-wave communication," *Elec. Commun.*, vol. 16, pp. 75–77; July, 1937. (Also *Electronics*, vol. 10, p. 52; December, 1937.

[3] S. Matsumae, S. Yonezana, and H. Kurokawa, "Multiplex carrier telephony on ultra-short wave at the Strait of Tugaru," *Nippon Elec. Commun. Eng.*, no. 20, pp. 206–219; April, 1940.

[4] H. S. Black, "Stabilized feedback amplifiers," *Bell Sys. Tech. Jour.*, vol. 13, pp. 1–18; January, 1934.

[5] C. W. Green and E. I. Green, "A carrier telephone system for toll cables, *Bell Sys. Tech. Jour.*, vol. 17, pp. 80–105; January, 1938; *Proc. A.I.E.E.*, vol. 57, pp. 227–236.

miles and to the Onancock central office 40 miles away is accomplished at voice frequencies.

Transmission from Cape Charles to East Ocean View is accomplished on a carrier frequency of 156,300 kilocycles and in the reverse direction on 160,650 kilocycles. This does not represent the minimum possible frequency spacing for this equipment, but was a convenient one which could be obtained.

Radio Equipment

The transmitters are crystal-controlled and have a carrier power output of 50 watts. The crystals are temperature-controlled in order to improve the frequency stability and thereby minimize the bandwidth and noise requirements at the receiver.

A block schematic of the transmitters is given in Fig. 2. The crystal oscillator in which the second har-

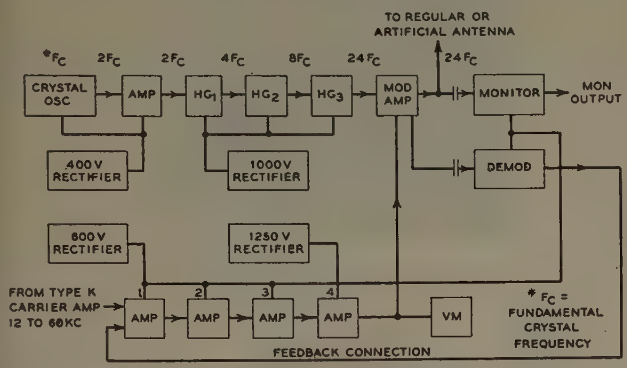

Fig. 2—Block schematic for D-159609 radio transmitter.

monic is selected is followed by a stage of amplification, three stages of harmonic generators, and a push-pull modulating amplifier. The final carrier frequency is the twenty-fourth harmonic of the crystal frequency. The amplitudes of the type K frequencies are increased by a 4-stage amplifier before they are applied to the modulating amplifier. The output of the transmitter consists of a carrier with sidebands extending from 12 to 60 kilocycles on each side of the carrier as shown in Fig. 3.

the tubes operate on alternating current and the plate, screen, and bias potentials are obtained from four single-phase mercury-vapor rectifiers. The transmitters are designed to operate into a balanced impedance of 140 ohms. A view of the transmitter is given in Fig. 4.

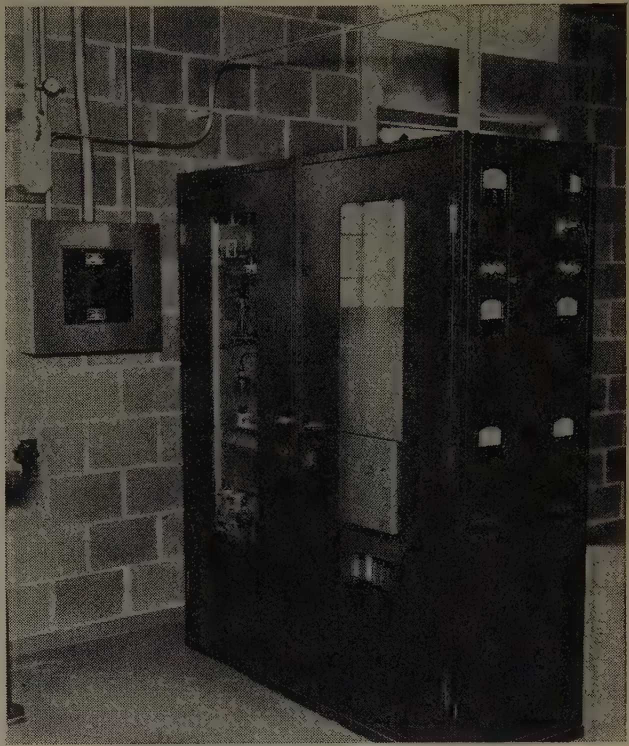

Fig. 4—D-159609 radio transmitter.

The input circuits of the receivers were designed to connect to a single 72-ohm coaxial transmission line. Triple detection was adopted in order to obtain sufficient image suppression and to satisfy other requirements. A block schematic of the receiver is given in Fig. 5. The receivers consist essentially of a first detector, a single-stage first-intermediate-frequency amplifier, a second detector, a 4-stage second-intermediate-

Fig. 3—Channel-sideband distribution.

The twelve channels thus take up a total band of 120 kilocycles or 10 kilocycles per channel. This amounts to a considerable saving in frequency space since current radio assignments in this region are on the basis of 150 kilocycles for a single channel. About 40 decibels of envelope feedback is obtained by detecting the output of the modulating amplifier by means of the demodulator and applying a portion of the detected output to the input of the 4-stage amplifier. A monitor which is similar to the demodulator is provided so that the character of the transmitter output can be observed without disturbing the circuits in any way. The filaments of all

frequency amplifier, and a high-level-diode final detector. The last three stages of the second-intermediate-frequency amplifier employ negative feedback in order to deliver a large signal input to the final detector with negligible distortion. Two additional amplifiers are bridged onto the circuit just ahead of the final detector. The output of one of these amplifiers is rectified and then used for automatic-volume-control purposes. The output of the other is passed through a sharp crystal filter, rectified and then used to operate a relay. The operation of this relay thus indicates that the frequency deviations of the distant transmitter and the local

receiver heterodyne oscillators have not displaced the second intermediate frequency more than ± 0.002 per cent of the radiated frequency.

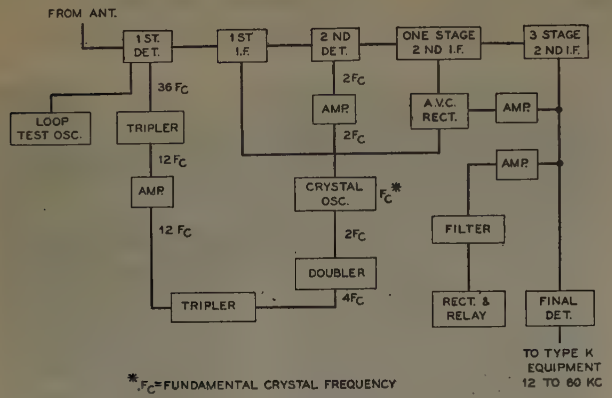

Fig. 5—Block schematic D-159608 radio receiver.

The heterodyne-oscillator supply for the first and second detectors is obtained from a single temperature-controlled crystal oscillator of the same type as is employed in the transmitter. The supply to the first detector is the 36th harmonic of the crystal and the supply to the second detector is the second harmonic. The first intermediate-amplifier frequency therefore depends upon the frequency to be received. The nominal band of the second intermediate frequency is from 1430 to 1570 kilocycles. A view of the receiver is given in Fig. 6.

Fig. 6—D-159608 radio receiver.

Separate, directive, horizontally polarized antennas are used for transmitting and receiving. The two antennas at each radio terminal are mounted one above the other on a single 196-foot steel lattice-type tower. The upper antenna at each end is used for transmitting and the lower one for receiving so as to equalize the transmission losses in each direction. A view of the Cape Charles tower and antenna is given in Fig. 7.

Fig. 7—Antenna installation.

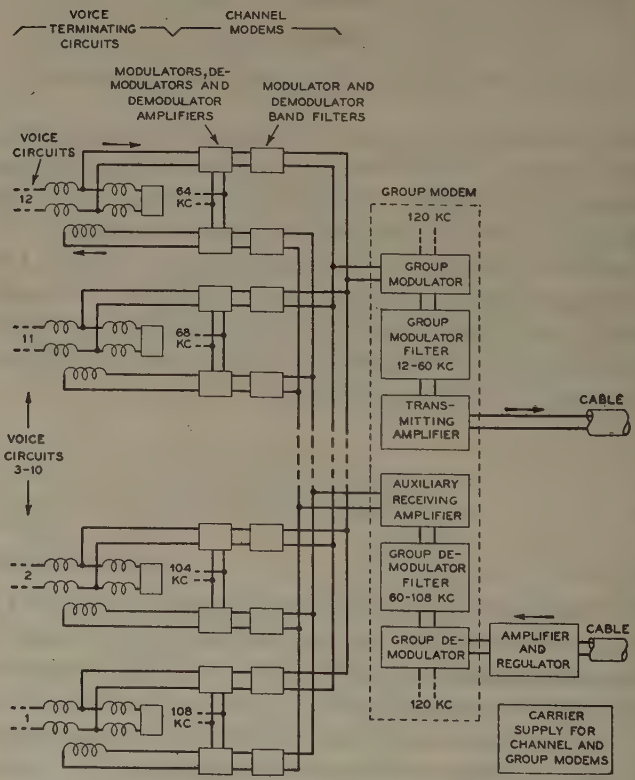

Fig. 8—Block diagram of type-K carrier terminal equipment.

The antennas consist of 48 one-half-wave elements arranged in two parallel vertical planes one quarter wave length apart. The elements in the plane farthest away from the tower are driven in phase whereas those

in the rear plane act as reflectors. Transmission is, of course, normal to these planes. The elements in both planes are made of one-half-inch-diameter copper tubing which are supported on a wooden framework by means of standoff insulators. The wooden framework for each antenna is in turn supported by six horizontal timbers which extend across the sides of the tower and are bolted to the legs of the tower. The upper antenna framework is somewhat larger than the lower in order that it can support four additional rods on the bottom and a long ground rod across the top. The upper rod is for lightning protection whereas the lower rods give additional attenuation between the two antennas. Each antenna has a gain of 17 decibels over a single half-wave element at the same mean height.

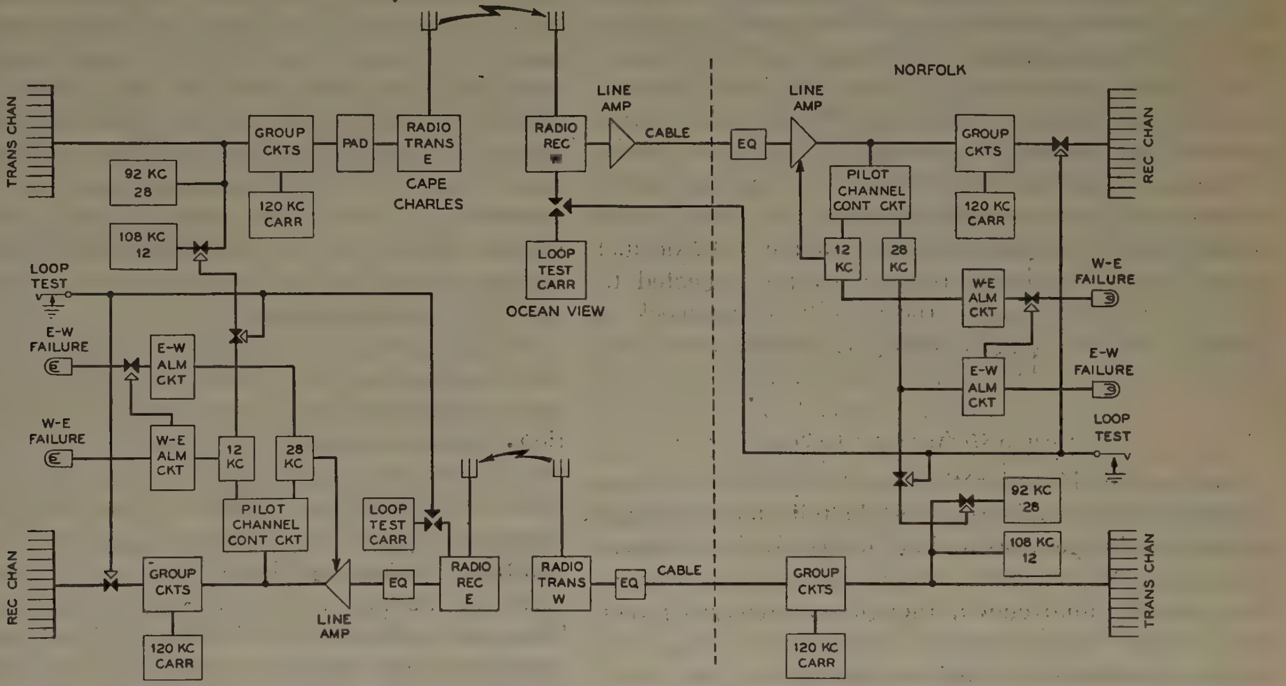

Fig. 9—Block diagram of the Cape Charles-Norfolk ultra-high-frequency multiplex system.

A pair of $\frac{7}{8}$-inch coaxial transmission lines connect the transmitter to its antenna while a single $\frac{7}{8}$-inch line is used in the case of the receiver. These lines are kept under gas pressure.

Carrier Terminal Equipment

In Fig. 8 is shown a block diagram of the type K carrier terminal equipment[6] used to place the 12 voice channels in the frequency spectrum from 12 to 60 kilocycles. This positioning is done in two steps for reasons of economy. A hybrid arrangement is associated with each voice circuit to transform the 2-wire circuit to a 4-wire circuit. The channel modems,[7] supplied with twelve carrier frequencies spaced 4 kilocycles apart from 64 to 108 kilocycles shift the voice channels to the range 60 to 108 kilocycles in the transmitting direction and restore the 60- to 108-kilocycle range to voice channels in the receiving direction. Associated with the channel modems are modulator and demodulator band filters for frequency selection and demodulator amplifiers to provide the desired volume to the voice circuits. The group modem, supplied with a 120-kilocycle carrier frequency shifts the 60- to 108-kilocycle band to the range of 12 to 60 kilocycles for application to the radio transmitter, and shifts the 12- to 60-kilocycle band from the radio receiver to the 60- to 108-kilocycle band utilized by the channel modems in the receiving direction. Associated with the group modem are two band filters and two amplifiers for frequency selection and gain. To minimize interchannel crosstalk the group modulator and demodulator are worked at low levels. This requires additional gain after each modulator and each demodulator.

Regulation

Although the radio receiver employs automatic volume control to overcome variations in loss over the radio path in accordance with normal radio practice, additional regulation is provided by the carrier equipment to maintain a substantially constant loss from terminal to terminal. The regulation is accomplished by the equipment shown schematically in Fig. 9.

Two pilot frequencies of the standard type K carrier are applied to the transmitting leg of the circuit just ahead of the group modulator. The frequencies, 108 and 92 kilocycles, are held to a constant amplitude by compensated thermistors and are shifted to 12 and 28 kilocycles by the 120-kilocycle group carrier for transmission over the cable and radio circuits to the carrier

[6] R. W. Chesnut, L. M. Ilgenfritz, and A. Kenner, "Cable carrier telephone terminals," *Bell Sys. Tech. Jour.*, vol. 7, pp. 106–124; January, 1938; *Proc. A.I.E.E.*, vol. 57, pp. 237–244.

[7] The term modem has been coined to mean a panel or equipment unit in which there are both a modulator and a demodulator to take care of both the outgoing and the incoming signals.

terminal at the far end. At the output of the line amplifier at the distant end, crystal filters select the two pilots and apply them to the pilot channel control circuit. The 12-kilocycle pilot is used to hold the circuit loss constant in the Cape Charles-Norfolk direction and the 28-kilocycle pilot is used for the same purpose in the opposite direction. The 28 kilocycles in the East-West[8] direction and 12 kilocycles in the West-East direction, are used for transmission alarms and in the loop tests, as described later.

The pilot regulation of the receiving-line amplifier gain is accomplished by standard type K carrier circuits. The pilot frequency of 12 or 28 kilocycles is selected by the appropriate filter, amplified and rectified to control the amplifier gain. The degree of regulation obtained with the circuit is such that for ± 11-decibel variation in input of the pilot the output of the pilot varies less than ± 0.5 decibel. Except in cases of trouble the variation in loss over the radio and cable circuits is considerably less than ± 11 decibels so that the variation in loss between the channel equipments of the two carrier terminals is less than ± 0.5 decibel. The variation in the channel equipment at each end does not exceed ± 0.5 decibel and of the pilot-frequency source ± 0.2 decibel so the over-all variation is normally not expected to exceed ± 0.8 decibel. For the circuits to Onancock, a further variation of ± 0.7 decibel is expected, resulting in a total of ± 1.1 decibels probable variation.

Control and Alarm Features

The system is under the direct supervision of the technical operators in the Norfolk toll office, who are licensed radio operators and are in attendance 24 hours a day. The Ocean View and Cape Charles radio stations are normally unattended, though maintenance forces are available when necessary. There is, of course, continual attendance at the switchboard in the Cape Charles central office.

A control, extended over cable circuits, enables the radio operator at Norfolk to shut off the Ocean View transmitter. By communication with the switchboard operator at Cape Charles, either over these circuits or via an overland wire circuit, orders to shut off the Cape Charles transmitter can be given. This transmitter power is controlled over wire circuits by a key mounted in the Cape Charles switchboard.

A continual check on the distant transmitter is supplied at both attended locations. As previously described, in each radio receiver is a sharply selective filter, tuned to the second intermediate frequency, and connected to a rectifier and relay. If the distant carrier is absent, or if the deviation of the transmitter and receiver beating oscillator is more than ± 0.002 per cent of the assigned value an alarm is given.

In addition to these features, which are designed to give to the technical operators at Norfolk effective control over the system, there are supplied transmission alarms and controls to facilitate maintenance. With a 12-channel system, it is highly desirable that the attendants at both ends of the circuit be notified promptly if the common path is opened in either direction. This is accomplished by the means shown schematically in Fig. 9. The 12-kilocycle pilot frequency sent over the circuit from Cape Charles is under control of the 12-kilocycle pilot received from Norfolk; if 12 kilocycles are not received from Norfolk, none is sent out toward Norfolk. Similarly, the 28-kilocycle pilot frequency sent from Norfolk is controlled by the 28-kilocycle pilot received from Cape Charles. If both frequencies are received at both locations, none of the alarms is operated. If a transmitter failure occurs in the Cape Charles-Norfolk direction (East-West) 28 kilocycles will not be received at Norfolk, and the E-W failure alarm will be operated. At the same time, the W-E failure alarm at Norfolk will be disabled and the 28-kilocycle pilot will be cut off in the other direction. This will cause operation of the E-W failure alarm at Cape Charles. Thus attendants at both circuit terminals will be notified of trouble. A failure in the other direction of transmission will operate similarly on the 12-kilocycle pilot, bringing in the W-E failure alarms.

It is a practical necessity to know on which side of the water barrier the failure has occurred, so that a maintenance man can be sent to the proper unattended location. This is accomplished by means of a near-end loop test. Under the conditions mentioned above operation of the loop test key at Norfolk restores the 28-kilocycle pilot frequency to the transmitting leg, and applies an additional beating oscillator frequency to the first detector of the Ocean View receiver, this being of such frequency as to tune in the Ocean View transmitter. Since the failure was at the other end of the circuit, both pilot frequencies will now be received, and consequently the E-W failure alarm will be released, indicating the failure to be in the transmitting leg at Cape Charles. Under these conditions if the loop test key is operated at Cape Charles, the local receiver will be tuned to the local transmitter, but the 28-kilocycle pilot will still be absent, thus the E-W failure alarm will still be operated, indicating also at this point a failure in the Cape Charles transmitting leg.

It will be noted that operation of the loop-test key also opens the circuit between the carrier channels and the group circuits, so as to prevent the possibility of singing around the near-end loop, across the hybrid coils which terminate the telephone channels.

To facilitate trouble location tests, jacks are provided in the line which is connected to the radio transmitter, at the output of the transmitter monitoring circuit and in the line which is connected to the radio receiver. Further, the circuit is arranged so that these are equal level points. With the loop test key operated, if trouble is indicated on the near side of the water a patch

[8] In normal telephone parlance all toll circuits are designated as if they run from East to West. By arbitrary convention, Cape Charles is the "East" and Norfolk the "West" terminal.

connection can be made from the transmitter monitoring circuit to the receiving line jacks. If the failure alarm is released, it is apparent that the trouble is in the radio receiver. If the alarm is not released a patch connection can be made between transmitting and receiving line localizing the trouble either in the radio transmitter or in the wire lines and carrier equipment.

Because of the fact that the pilot-channel control circuit and the associated line amplifier do not compensate for changes of loss instantaneously it is necessary to make the E-W and W-E alarm circuits relatively slow in operation. If this were not done a sudden decrease in loss of a few decibels would bring in a temporary false alarm. Slow operation of the alarm circuits is obtained by the use of temperature-compensated thermistors which give slower operation of the alarm circuits than the regulation time of the controlled-line amplifier.

At the Norfolk end the E-W and W-E alarms appear only at the toll office since the carrier equipment is located there. At Cape Charles they appear both at the radio station and the central office.

Other Alarms

At the western end of the circuit alarms are provided both at Ocean View and the Norfolk Toll Office for an open radio-station door, high radio-cabinet temperature, alternating-current power failure, low or high radio-station room temperature, and fire. At Norfolk the first three are grouped together to light a Major Alarm lamp, the fourth lights a Minor Alarm lamp, and the fifth lights a Fire Alarm lamp. Two pairs in the cable between Ocean View and Norfolk Toll Office are required for the transmission of these alarms. An additional pair permits performance of the loop test and shutdown of the radio transmitter from Norfolk.

At the eastern end of the circuit, alarms are provided both at the radio station and the Cape Charles Central Office for open radio-station door, high radio-cabinet temperature, low or high radio-station room temperature, fire, and various power-supply failures. As at Norfolk some of these were grouped into major and minor alarms for transmission to the central office. Four cable pairs with ground return are required for transmission of these alarms. An additional pair permits performance of the loop test and shut down of the radio transmitter from the central office.

Systems Tests

Intermodulation tests using both tones and speech have been made from terminal to terminal of the complete system. In each test the signal was applied to one or more channels and the intermodulation measured in the remaining channels. The results of the more significant speech tests are given in Fig. 10. Curve 1 represents

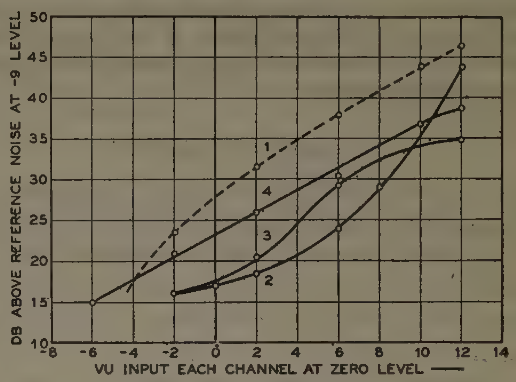

Fig. 10—Performance curves.

the limit below which it was desirable to keep the intermodulation based on the assumption that the link should be at least as good in performance as 1000 miles of type K carrier equipment in summer. The abscissas give the input to each channel at the zero level point in the radio circuit in VU. The ordinates give the noise in decibels above reference noise at −9 level as measured with the 2B noise meter. Curve 2 represents third-order modulation in channel 10 resulting from speech in channels 11 and 12. Curve 3 represents second-harmonic modulation in channel 12 resulting from speech in channel 5. Curve 4 represents second-order modulation in channel 1 resulting from speech in channels 9 and 12.

As can be seen the results are well below the design limit. With these gain adjustments 1 milliwatt of testing power at zero level in one channel gives 8 per cent modulation.

This new link was placed in service in October, 1941, with two of the channels used for traffic between Norfolk and Cape Charles and three others used between Norfolk and the toll center at Onancock. All channels are now in service, five between Norfolk and Cape Charles, four between Norfolk and Onancock and three for "private-line" telephone circuits.

Correction

Major Edgar H. Felix has drawn to the attention of the editors the following corrections to his paper, "The Use of Field-Intensity Measurements for Commercial-Coverage Evaluation," which appeared on pages 381 to 393 of the July, 1944, issue of the Proceedings: In the charts which appeared on pages 387, 388, and 389, the word "microvolts" should be changed to "millivolts."

Ultra-Short-Wave Multiplex*

CHARLES R. BURROWS†, FELLOW, I.R.E., AND ALFRED DECINO‡, SENIOR MEMBER, I.R.E.

***Summary*—The technical requirements of a twelve-channel ultra-short-wave multiplex system are discussed and the means of meeting them are described. The intermodulation between channels in equipment based on this design has been reduced to the point where it is possible to use twelve-channel radio systems in the toll plant. By employing a sufficient amount of envelope feedback, the transmitter can be operated with a high modulation factor without the use of spread sidebands.**

INTRODUCTION

THE advantages possessed by multiplex in wire telephony obviously call for the evaluation of corresponding methods in radio communication. Over a period of years, different phases of this problem have been studied in the Bell Telephone Laboratories by various investigators. The present paper is one of a group of four companion papers which discuss the development of equipment suitable for the transmission of twelve telephone channels over radio spans in the telephone plant. The radio can be used as part of a very much longer wire system. This paper deals primarily with the circuit design of the transmitter modulating arrangements.

This system employs the twelve telephone channels of the type-K cable-carrier system which are in the frequency range 12 to 60 kilocycles per second as the signal to modulate a radio-frequency carrier of about 160 megacycles per second. The development during 1940 was specifically aimed at the transmission of the output of a type-K twelve-channel sending terminal[1,2] across the mouth of Chesapeake Bay. This system is described in detail in a companion paper.[3] Obviously, however, the principles to be studied are of broader application.

When considering the noise permissible in a multiplex telephone circuit, a very severe requirement is encountered. The average noise that may be introduced into one channel by the whole toll circuit is commonly expressed as 60 to 70 decibels below the peak load capacity required by the composite multiplex signal at any point. This must cover both the many varieties of "noise" and the intermodulation products produced in the channel under observation by intermodulation from all other channels. The amount of intermodulation may also be determined by sending through one test tone or a pair of test tones, and measuring a single frequency-modulation product. In this form of test, the maximum permissible amplitude of the intermodulation product is about 40 decibels below the single test tone when its amplitude corresponds to the load capacity of the composite multiplex signal. This requires a fidelity of transmission that is unusual in radio transmitters. The requirement might be met by reducing the sideband power to a small fraction of the nominal power capacity of the transmitter but as this method also decreases the signal-to-noise ratio at the receiver, antennas with more gain, or higher antenna towers would be needed, or the system would be limited to a shorter transmission span. Economical design requires the reduction of the intermodulation to a satisfactory value at a sideband power approaching the nominal power capacity of the transmitter.

There are several ways of meeting this distortion requirement. In the nine-channel multiplex system (80 megacycles per second) between Scotland and Ireland across St. George's Channel[4] the distortion problem is mitigated by using the spread-sideband system. In that system there is a frequency range on either side of the carrier which is equal to or greater than the width of the sideband so that the predominant distortion products fall outside of the useful band. This system doubles the frequency range required and makes it greater than the frequency interval between assignments in this country. Furthermore, the British employ a double-sideband signal for each individual channel as well, which multiplies the required frequency range by another factor of 2. The Japanese have a six-channel 75-megacycle multiplex system[5–7] bridging the Straits of Tugaru. It employs a type of negative feedback to reduce the crosstalk but even with the amount of stable feedback obtained (about 15 decibels) it was necessary to employ the spread-band system to reduce the crosstalk to a satisfactory value.

For present purposes it was considered undesirable to obtain low intermodulation at the cost of operation either at low modulation levels or with spread sidebands. Single-sideband transmission was considered, but the special complexities involved in the terminal problem and the difficulties in obtaining adequate feedback in

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[1] C. W. Green and E. I. Green, "A carrier telephone system for toll cables," *Bell Sys. Tech. Jour.*, vol. 17, pp. 80–105; January, 1938.

[2] R. W. Chesnut, L. M. Ilgenfritz, and A. Kenner, "Cable carrier telephone terminals," *Bell Sys. Tech. Jour.*, vol. 17, pp. 106–124; January, 1938.

[3] N. F. Schlaack and A. C. Dickieson, "Cape Charles-Norfolk ultra-short-wave multiplex system," PROC. I.R.E., this issue, pp. 78–83.

[4] E. H. Ullrich, "Ultra-short-wave communication," *Elec. Commun.*, vol. 16, pp. 64–86, July, 1937. (See also *Electronics*, vol. 10, p. 52; December, 1937.)

[5] S. Yonezawa and Y. Hirayama, "Relation between non-linear distortion and crosstalk on multiplex transmission," *Nippon Elec. Commun., Eng.*, no. 11, pp. 215–230; June, 1938.

[6] S. Matsumae and S. Yonezawa, "Equipment for multiplex carrier telephony on ultra-short wave," *Nippon Elec. Commun. Eng.*, no. 15, pp. 554–563; February, 1939.

[7] S. Matsumae, S. Yonezawa, and H. Kurokawa, "Multiplex carrier telephony on ultra-short wave at the Strait of Tsugam," *Nippon Elec. Commun. Eng.*, no. 20, pp. 206–219; April, 1940.

this frequency range[8] as compared with the problem if the double-sideband method were used, led to the adoption of the latter. A survey of tubes, at that time available, indicated that more undistorted sideband power could be obtained from a modulated radio-frequency amplifier producing a double-sideband system than from a single-sideband radio-frequency amplifier. The double-sideband system also allowed the use of envelope feedback which greatly simplifies the feedback problem. In order to obtain the desired signal-to-distortion ratio with a double-sideband system, however, the departures of the phase characteristic of the receiver from skew symmetry with respect to the carrier frequency must be considerably less than is usual in the general run of receivers. After considering the various phases of the problem, it was decided to build a double-sideband system with envelope feedback.

DISTORTION REQUIREMENTS

In employing envelope feedback, a double-sideband multiplex system, interchannel modulation may be caused in several ways. These may be classified under three general types: (1) nonlinear distortion, (2) deviation from the ideal of the over-all gain-phase-frequency characteristic of that part of the circuit between the modulator at the transmitter and the final demodulator at the receiver, and (3) generation and detection of parasitic phase modulation. Topics to be discussed are:

- I. Nonlinear distortion
  - A. Transmitter
    - 1. Signal-frequency amplifier
    - 2. Modulator
    - 3. Demodulator in feedback loop
  - B. Receiver
    - 1. Intermediate-frequency amplifier
    - 2. Demodulator
    - 3. Signal-frequency amplifier
- II. Nonideal gain-phase-frequency characteristic
- III. Generation and detection of parasitic modulation

I. Nonlinear Distortion

Any deviation from linearity of the input-output amplitude characteristic will generate new frequencies causing intermodulation. This may occur in the signal-frequency amplifier either at the transmitter or the receiver or in the modulator at the transmitter or the demodulator at the receiver.

A. Transmitter

When high-level modulation is employed, it is usually the modulator which produces the predominant nonlinear distortion. By proper design, the radio-frequency output is made as nearly proportional to the plate voltage on the modulator as possible, but for any given modulation factor, there is an irreducible minimum beyond which the distortion cannot be reduced without the use of feedback. In order to reduce this distortion sufficiently, it is necessary to employ a larger amount of feedback than has hitherto been employed in an envelope-feedback system of this bandwidth. The method of obtaining this feedback is described in a later section on Feedback Design.

The use of a large amount of envelope feedback, however, alone cannot solve the problem of nonlinear distortion, because envelope feedback requires an active element in the β circuit, the demodulator. Any distortion generated in the demodulator is, accordingly, present in the output of the transmitter. The demodulator may be a high-level linear detector whose distortion is inherently less than that of the modulator so that there is some improvement from feedback, but the amount of improvement is limited by the linearity of the demodulator and not by the amount of feedback. The nonlinear distortion problem at the transmitter thus involves the design of a demodulator whose distortion is very low, as well as the provision of a sufficient amount of envelope feedback.

B. Receiver

The demodulator is likely to be the predominant source of nonlinear distortion in the receiver. Distortion from this cause may be minimized by employing a high-level linear detector. To obtain the necessary high level to drive this type of demodulator, negative feedback is used to suppress the generation of nonlinear distortion in the intermediate-frequency amplifier.

Another method of reducing the distortion produced by the demodulator in the transmitter and the demodulator of the receiver is to make them the same, so that the distortion produced by the demodulator in the transmitter is canceled by a like distortion of opposite sign in the demodulator of the receiver. The over-all nonlinear distortion will then be less than that from either demodulator. With sufficient feedback the amount of reduction of the distortion below that of either demodulator is a measure of the equivalence of the two demodulators. The practical difficulty of obtaining and maintaining the equivalence of these two demodulators makes it appear desirable to reduce the distortion generated in each without regard to that generated in the other.

II. Nonideal Gain-Phase-Frequency Characteristic

A double-sideband multiplex system places severe requirements on the transmission characteristic from the modulator to the demodulator. The requirement is that the complex gain at any frequency on one side of the carrier be the conjugate of that at the corresponding frequency on the other side when both gains are relative to that at the carrier frequency. This results in a gain characteristic that is symmetrical with respect to the

[8] Some years ago, R. C. Shaw of these Laboratories, developed an amplifier having about 15 decibels of feedback at 125 megacycles. More recent considerations significantly augment the amount that can be obtained with present tubes and methods. Such an amplifier makes faithful amplification of single-sideband multiplex feasible.

carrier frequency, and a phase characteristic that is skew symmetrical. Skew symmetry means that the phase at any frequency on one side of the carrier is the negative of that at the corresponding frequency on the other side of the carrier when both phases are expressed as relative to that at the carrier. Any deviation from this type of symmetry introduces new frequencies. The way in which this occurs is shown in the Appendix. The amplitudes of the predominant second-order distortion products for two test tones are found to be

$$(m_p^2/16)(\theta_p^2+\delta_p^2) \quad \text{and} \quad (m_q^2/16)(\theta_q^2+\delta_q^2) \tag{1}$$

for the second harmonics, and

$$(m_p m_q/8)\sqrt{(\theta_p^2+\delta_p^2)(\theta_q^2+\delta_q^2)} \tag{2}$$

for the sum and difference frequencies. Here m_p and m_q are the modulation factors and θ_p and θ_q are the deviations of the phase characteristic from skew symmetry in radians for the sidebands in question. Each δ is the difference of the ratios of the gains at the two side frequencies in question to that at the carrier frequency. These distortion products should be compared with the desired signal amplitudes which in the units used in (2) are equal to the modulation factors m_p and m_q. In making distortion tests it is customary to use equal modulation factors for both test tones and to express the distortion products relative to either signal amplitude. When this is done, these distortion-to-signal ratios become

$$(m/16)(\theta_p^2+\delta_p^2), \quad (m/16)(\theta_q^2+\delta_q^2) \quad \text{and}$$
$$(m/8)\sqrt{(\theta_p^2+\delta_p^2)(\theta_q^2+\delta_q^2)}. \tag{3}$$

III. Generation and Detection of Phase Modulation

Envelope feedback does not affect any phase modulation that may be inadvertently produced in the modulation process. A linear detector responds only to the envelope of the carrier and does not detect any variations in its phase or frequency. But deviations of the transmission characteristic from symmetry may change the carrier envelope even if there is no phase modulation present as pointed out in the previous section. With phase modulation present in the transmitter additional requirements must be met.

For an ideal receiver that has a constant gain and linear phase within the band but a zero response outside the band, the second-harmonic-to-fundamental ratio is, to a first approximation,

$$\Phi^4/48m, \tag{4}$$

where m is the amplitude modulation factor and Φ is the phase-modulation index. This assumes the worst condition in which the second-harmonic distortion lies within the transmitted band but the fourth harmonic does not. If the fourth harmonic is also transmitted, the second harmonic distortion is reduced. Conditions leading to (4) do not put a stringent requirement on the phase modulation generated by the transmitter. To obtain a fundamental-to-second harmonic ratio of 100 (40 decibels) at an amplitude modulation of 0.8 (80 per cent) would require that the phase-modulation index associated with this amplitude modulation factor should not be more than 0.783 radian or ±45 degrees. A transmitter that meets this requirement for large modulation factors would be expected to meet it for smaller factors since the parasitic phase modulation would decrease with the amplitude modulation associated with it.

When the receiver characteristic deviates from the above ideal, more terms are added to the distortion. W. T. Wintringham has calculated the second-harmonic distortion under these conditions when the bandwidth is just enough to pass this distortion. His results for the second-harmonic-to-fundamental distortion ratio may be expressed in the form

$$
\begin{aligned}
&(1/8m)\{(\Phi^4/6)^2\\
&+(1/3)\Phi^6\delta_l+\Phi^4(\delta_l^2\ \theta_l^2)\\
&+(2/3)m\Phi^5\theta_s+4m\Phi^3(\delta_l\theta_s-\theta_l\delta_s)-(1/6)m^2\Phi^4(\delta_0^2-\theta_0^2)\\
&+m^2\Phi^2[\delta_l(\delta_0^2-\theta_0^2)+4\theta_s^2+2\theta_l\delta_0\theta_0+4\delta_s^2]\\
&\quad-2m^3\Phi[\theta_s(\delta_0^2-\theta_0^2)+2\delta_0\delta_s\theta_0]\\
&+(m^2/2)^2(\delta_0^2+\theta_0^2)^2\}^{1/2}
\end{aligned} \tag{5}
$$

where

$$
\begin{aligned}
m &= \text{amplitude-modulation factor}\\
\Phi &= \text{phase-modulation index}\\
\delta_l &= 2(A_p+A_{-p})-(A_{2p}+A_{-2p})-2\\
\delta_s &= (A_p-A_{-p})-(A_{2p}-A_{-2p})\\
\delta_0 &= (A_p-A_{-p})\\
\theta_l &= 2(\phi_p-\phi_{-p})-(\phi_{2p}-\phi_{-2p})\\
\theta_s &= (\phi_p+\phi_{-p})-(\phi_{2p}+\phi_{-2p})\\
\theta_0 &= (\phi_p+\phi_{-p})
\end{aligned} \tag{6}
$$

and the gain-phase-frequency characteristic is proportional to

$$A_x \exp i(k_1\omega+k_2x+\phi_x) \tag{7}$$

at any frequency $\omega+x$. Here ω is the carrier frequency, k_1 and k_2 are constants and ϕ_x is the departure of the phase from linearity. The implied proportionality factor is adjusted so that $A_0=1$, and k_1 is chosen so that $\phi_0=0$. The first line of (5) gives the term that would be present even if the receiver were ideal, the second line gives the additional term if the receiver characteristics have the proper symmetry, and the third, fourth, and fifth lines give the additional terms if the receiver characteristics lack symmetry. The sixth line gives the distortion ratio that is present even in the absence of phase modulation.

The first term is small even for excessive amounts of phase modulation. The terms in the second line do not depend upon symmetry but tend to be negligibly small unless there are large ripples in the gain and phase characteristics. Even if the ripples all have their maximum and minimum values in the most undesirable locations, a phase modulation index of 0.2 and ripples of ±1.3 decibels in gain and 9 degrees in phase will allow 57-decibel signal-to-distortion ratio provided the desired symmetry is maintained so that the remaining terms of (5) are zero. The greater importance of symmetry is evident from the last term which indicates that unsymmetrical deviations of the same magnitudes will

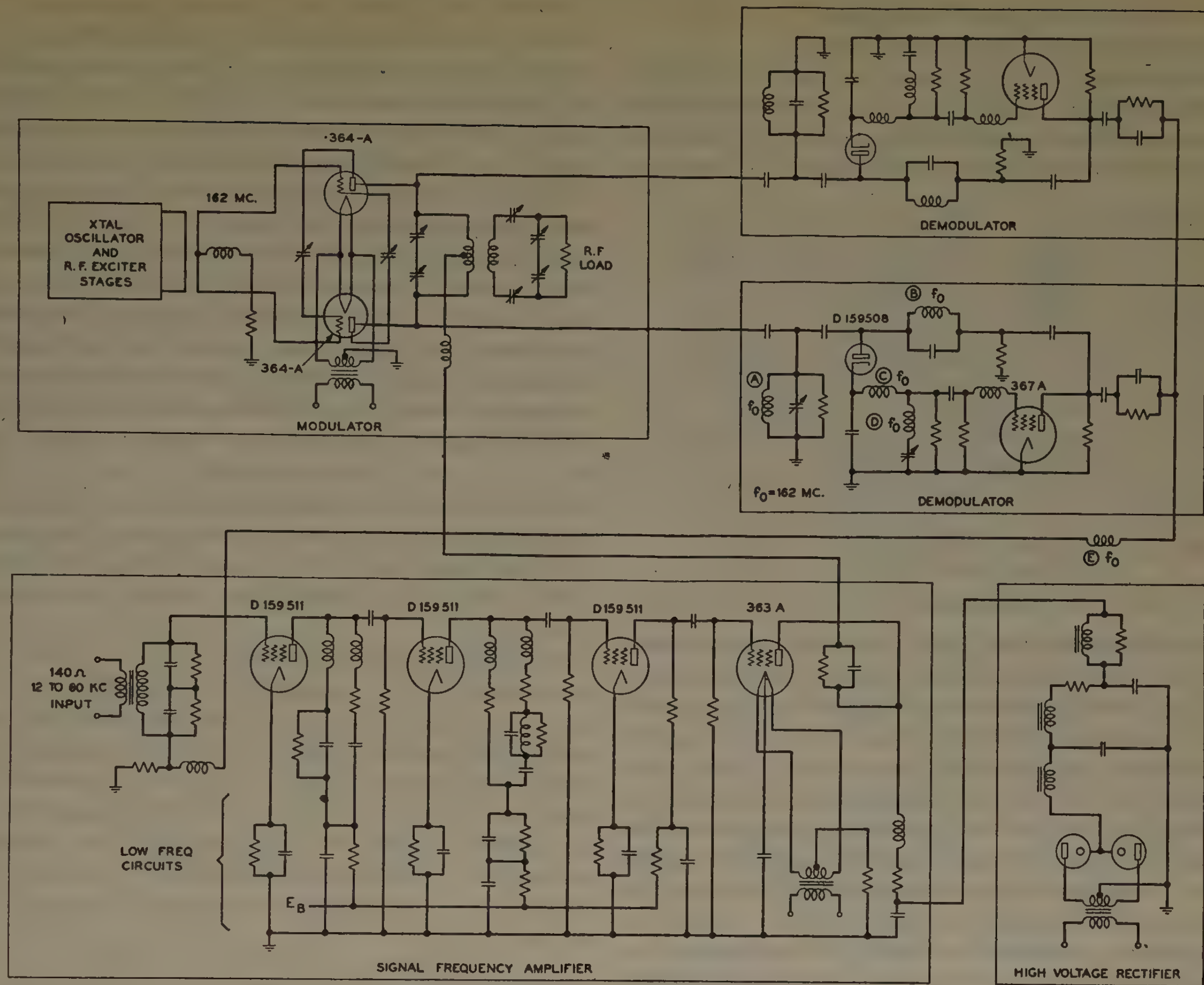

Fig. 1—Circuit schematic of the transmitter.

result in a 40-decibel signal-to-distortion ratio even if there is no phase modulation.[9]

TRANSMITTER DESIGN

Fig. 1 shows the circuit schematic of the transmitter designed to meet these requirements. This transmitter is described in detail in a companion paper.[10] The radio-frequency drive is obtained from a crystal oscillator and harmonic-generator chain. This is applied to the grids of two bridge neutralized 364A tubes operating as a push-pull modulator. The 12-speech channels in the frequency range from 12 to 60 kilocycles per second come from the type-K equipment at an impedance level of 140 ohms. This is applied to the grid of the first signal-frequency amplifier through a 140- to 30,000-ohm transformer. The first three stages employ new high-transconductance tubes. The last tube is a 350-watt pentode whose output plate modulates the 364A's.

Part of the output of the transmitter is demodulated and fed back to the input of the signal-frequency amplifier to provide negative feedback for distortion correction.

I. Feedback Design

In designing the feedback networks for commercial equipment, it is necessary to include margins to permit a reasonable tolerance in tube manufacture and a reasonable amount of aging of tubes before it becomes necessary to replace them.

Measurements made on the experimental model of the equipment indicate that 28 decibels of feedback is required to limit the nonlinear distortion to the desired value. The commercial equipment is designed for 38 decibels of feedback to allow for the decrease in transconductance as the tubes age. Variations in interelectrode capacitances and transconductances of different tubes will change the $\mu\beta$ characteristic. This requires a design that provides stability against these variations. In order to allow for this and to allow for a reasonable departure between the designed and realized characteristic, a phase margin of 30 degrees and a gain margin of 10 decibels are assumed in the design. This means that the amplifier will remain stable as long as the phase does not increase as much as 30 degrees or the gain as

[9] Numerical values in this paragraph refer to $m=0.8$.

[10] R. J. Kircher and R. W. Friis, "Ultra-high-frequency transmitter for the Cape Charles-Norfolk multiplex system," Proc. I.R.E., this issue, pp. 101–106.

much as 10 decibels from the design. No amount of decrease in gain will produce instability.

The realization of the necessary amount of feedback was made possible by two factors: (1) the development of new tubes which permit a larger amount of feedback and (2) the feedback design theory of H. W. Bode.[11]

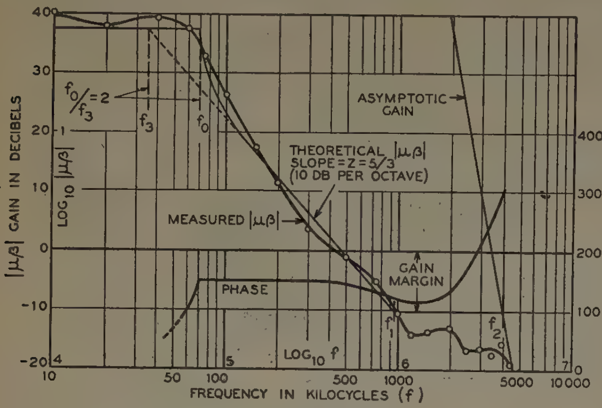

Fig. 2—The $\mu\beta$ characteristic of the transmitter.

In a radio transmitter, high transconductance is more important than low input and output capacitances because the tubes that have the higher figure of merit must be associated with higher-power tubes that inherently have a higher input capacitance. To meet this need the Bell Telephone Laboratories developed the D-159511 tube which has a transconductance of 10,000 micromhos with input and output capacitances of 10 and 6 micromicrofarads, respectively. This tube has sufficient power capacity to drive the 363A. The latter is a 350-watt pentode also developed particularly for this application. It has a transconductance of 12,000 micromhos with input and output capacitances of 29 and 21 micromicrofarads, respectively. When operated with a slightly positive voltage on the suppressor, this tube will deliver 65 watts into a load of 7000 ohms with second harmonic distortion 40 decibels below the signal. This tube is thus capable of driving the modulator without producing as much distortion as is inherent in the modulator.

According to Bode's design, the maximum possible amount of feedback is dependent upon the asymptotic $\mu\beta$ gain. This is defined as the total gain of the tubes operating into the parasitic capacitance associated with their plate circuits times the asymptotic β circuit loss. The asymptotic β circuit loss is determined by the capacitance network that results from ignoring other circuit elements. The loss of this capacitance potentiometer is made as small as possible consistent with other requirements. For the Heising system of modulation the modulator has the same effect as an additional stage with a mid-band gain equal to the ratio of signal voltage on the plate of the modulator to the signal voltage on the demodulator, and a bandwidth equal to half the bandwidth of the modulator plate circuit.

[11] H. W. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421–454; July, 1940.

Bode's design consists of shaping the $\mu\beta$ characteristic so that there is a constant gain margin (fixed loss) for frequencies remote from the band where the phase is unfavorable and a constant phase margin against parasitic oscillations near the band where the $\mu\beta$ gain is equal to or greater than unity. In Fig. 2 there is a gain margin of 10 decibels in the frequency interval f_2 to f_1, and a phase margin of 30 degrees in the interval f_1 to f_0 which depends upon the slope in this interval. Bode has shown that the most favorable phase margins result from minimum phase networks and that for this type of network the phase margin can be maintained by making $f_2/f_1=n/z$, where n is the asymptotic slope of the log gain—log frequency characteristic, and z is the slope required to give the desired phase margin, and

$$z = 2(1-\theta/\pi) \tag{8}$$

where θ is the phase margin in radians. In the present amplifier n is equal to the number of stages counting in the modulator ($n=5$). The present design is based on $\theta=\pi/6$ (30 degrees) which gives $z=5/3$ or 10 decibels per octave.

In actual amplifiers, however, there is a phase shift in addition to the minimum phase shift which is important except at low frequencies where transit times are negligibly small compared with the period of all the frequencies involved and the apparatus is small compared with the wavelength. The feedthrough due to grid-plate capacitance also introduces an additional phase shift. When these additional phase shifts are taken into consideration the foregoing equation is replaced by

$$f_2/f_1 = n/z + \pi^2 T f_2/z + \pi^2 f_2/f_s z \tag{9}$$

where T is the time delay due to the transit time of the tubes and the additional delay of any all-pass network that may be inherent in the amplifier, and f_s is the frequency at which the circuit is, in effect, 1 wavelength long. The last term accounts for the added delay resulting from the physical size of the feedback loop. As expressed, it gives the delay of a terminated transmission line one wavelength long at the frequency f_s. For design purposes it seems to be satisfactory to measure the over-all distance around the $\mu\beta$ loop for determination of f_s. Because the $\mu\beta$ gain is constant in the band, it is possible to increase the slope of the $\mu\beta$ characteristic near the edge of the band as shown on Fig. 2 without decreasing the phase margin. This fact allows the frequency of the upper edge of the useful band to be twice what it would have been if the constant slope were continued as shown by the dotted line in Fig. 2.

The realization of the desired $\mu\beta$ frequency characteristic with the simplest types of networks is an art. The feedback networks of the experimental models are the result of a design by R. L. Dietzold and W. H. Boghosian. In this particular case the β circuit is fixed by other considerations. The modulator is a radio-frequency network, so as simple a circuit as practical is used in it. It also is the load impedance for the last stage in the signal amplifier and therefore this stage is not

available for use in shaping the $\mu\beta$ curve. In order to operate the three earlier stages at low levels to reduce the distortion, the gain of the penultimate stage is made high. Since little shaping can be accomplished in a high-gain stage it is a simple resistance-coupled stage and all the $\mu\beta$ shaping is realized in the first two stages. On this basis, the next step is to calculate the contribution to the $\mu\beta$ characteristic of the remainder of the $\mu\beta$ loop and subtract it from the desired $\mu\beta$ curve. This result gives the characteristic that must be satisfied by stages one and two and is shown in Fig. 3.

The characteristic shown in Fig. 3 consists of the sharp corner at the upper edge of the band, a broad minimum followed by a peak at higher frequencies where the gain of the remainder of the $\mu\beta$ loop is falling rapidly but the desired characteristic has a constant gain equal to the gain margin. The networks for interstages 1 and 2 and their impedance frequency characteristics are shown in Figs. 4 and 5, respectively. The capacitance C_1 represents the total parasitic capacitance. The resistance R_1 together with the transconductance of the tube determines the gain within the band. The inductance L_1 is put in series with R_1 to enable a high impedance to be obtained at several megacycles. The capacitance C_3 is added to introduce the steep slope required just beyond the edge of the useful band. At low frequencies the impedance in series with C_3 becomes small compared with that of C_3. Accordingly, the network neglecting that impedance is used as the basis of design at frequencies near the edge of the band. Here the response of stage 1 has a minimum followed by a maximum, while that of stage 2 has a single maximum. By properly proportioning the circuit elements it is possible to obtain a characteristic which is substantially flat within the band and has approximately the steep slope required by the design. The remainder of the frequency characteristic must be realized by properly adjusting the circuit elements ignored until now.

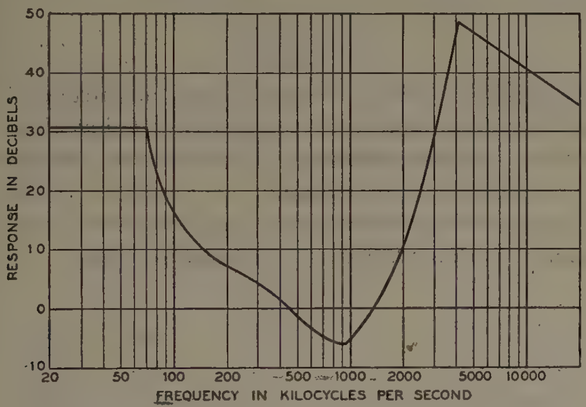

Fig. 3—Required gain-frequency characteristic of stages 1 and 2.

The series circuit ($L_2R_2C_3[L_2+L_3R_2C_3]$ for stage 2) controls the minimum at somewhat higher frequencies and in connection with the parasitic capacitance C_1 produces a peak at still higher frequencies. L_2 may be chosen so as to locate the frequencies of antiresonance or series resonance, but the capacitances C_1 and C_3 are already fixed so that the ratio of resonance to antiresonance is determined. Likewise R_2 may be chosen so as to determine the impedance at resonance or antiresonance. In the design under discussion it was found necessary to introduce the antiresonant circuit $L_3R_3C_2$ in series with $L_2R_2C_3$ in stage 2 in order to obtain the desired peak at 4.5 megacycles.

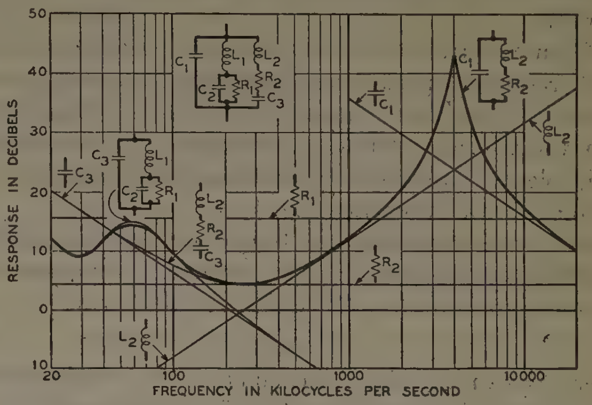

Fig. 4—First interstage network and gain-frequency characteristics.

In Figs. 4 and 5 the impedance characteristic of some of the component parts of the networks are also shown to indicate how they control the characteristic in various frequency ranges. These facilitate sketching in the network response and indicate the way the network should be changed to approach more nearly the desired characteristic.

The low-frequency $\mu\beta$ characteristic below the useful band is designed to satisfy similar phase and gain requirements except that now the gain-frequency slope is positive.

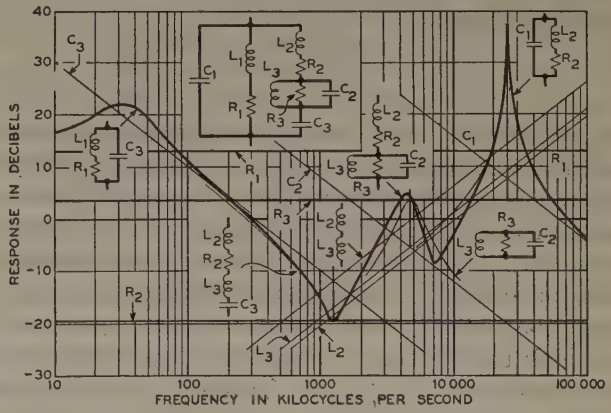

Fig. 5—Second interstage network and gain-frequency characteristics.

The amount of feedback for which the low-frequency characteristic may be designed is not restricted by the figure of merit of the tubes nor parasitic capacitances of the network but depends only upon the size of the capacitances and resistances of the network. Accordingly it is desirable to design this characteristic with greater margins in order to permit greater manufacturing tolerances in the network elements.

The size of the coupling capacitance is limited by the fact that larger capacitance means larger physical size and larger parasitic capacitance resulting in a lower high-frequency asymptotic gain for the stage. This affects the design of the high-frequency $\mu\beta$ characteristic so as to reduce the amount of feedback possible. The value of the grid resistance is limited by the grid current.

If a modulation choke is employed, the plate-supply filter requires special consideration. Unless its resonances and antiresonances are at such low frequencies that the $\mu\beta$ characteristic is already well cut off, it is desirable to damp these resonances by including resistances in the filter. See Fig. 1.

All of the remaining low-frequency network components have frequency characteristics composed of products and quotients of factors of the form

$$1 + \omega_a/p \qquad (10)$$

where $p = i\omega$ is as usual the coefficient of t in the assumed exponential time factor, and ω_a is a characteristic frequency of the network. The design problem is to build an over-all network that has a resultant slope of 9 decibels per octave out of factors of this type, each of which changes from a zero slope to a slope of 6 decibels per octave gradually over a two-octave range. (At the characteristic frequency the two asymptotes intersect and the curve is 3 decibels from either. At an octave either way from the characteristic frequency the curve is 1 decibel from its asymptote.)

This design may be accomplished by introducing two factors in the denominator of the expression for the gain with the same characteristic frequency and following them by factors alternately in the numerator and denominator whose characteristic frequencies are separated by a fixed ratio. This results in the following gain-phase-frequency characteristic

$$\frac{1}{(1 + \omega_0/p)^2} \prod_{s=0}^{\infty} \frac{(1 + \omega_0/n^{(2s+1)}p)}{(1 + \omega_0/n^{(2s+2)}p)}. \qquad (11)$$

Here n is the frequency ratio between the characteristic frequencies of the successive factors. Fig. 6 shows the gain-frequency characteristic for n equal to 2 and 4. Values of n very much greater than these will produce undesirably large undulations in the characteristic. In order not to reduce the amount of feedback appreciably within the useful band, ω_0 should be made equal to about half the frequency of the lower edge of the band.

It is not necessary that the frequency ratios between successive factors be all the same but only that the characteristic frequency of each factor in the denominator be the geometric mean of the adjacent factors in the numerator. This fact is useful in designing cathode networks because the ratio between the two factors of each cathode network is determined by the tube characteristic.

In order to reduce the impedance required of the modulation choke as much as possible, it is designed so that its impedance at the first characteristic frequency

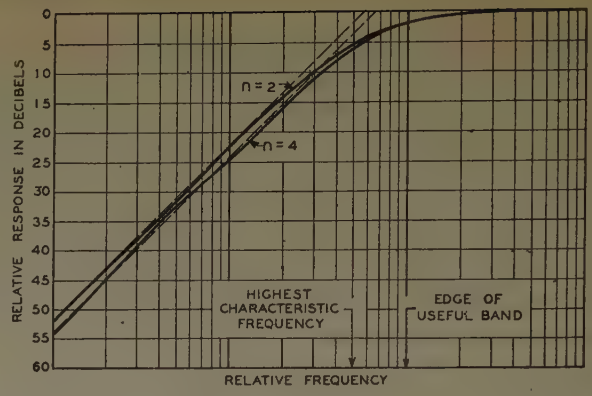

Fig. 6—Low-frequency design gain-frequency characteristics.

is equal to the mid-band circuit impedance. This characteristic frequency is made equal to about half that of the lower edge of the useful band. Another factor with the same characteristic frequency is then introduced either by a cathode network or by a coupling circuit.

Each cathode network of the type shown in Fig. 7a introduces the quotient of two factors, thus,

$$\frac{1 + \omega_2/p}{1 + \omega_1/p} \qquad (12)$$

where $\omega_1/\omega_2 = 1 + RG_m$ is equal to the magnitude of the local feedback introduced at frequencies below those for which the cathode resistance R_2 is effectively by-passed by the cathode filter capacitance C_2. Here G_m is the transconductance of the tube. The lower characteristic frequency is determined by the cathode filter.

$$\omega_2 = 1/R_2C_2. \qquad (13)$$

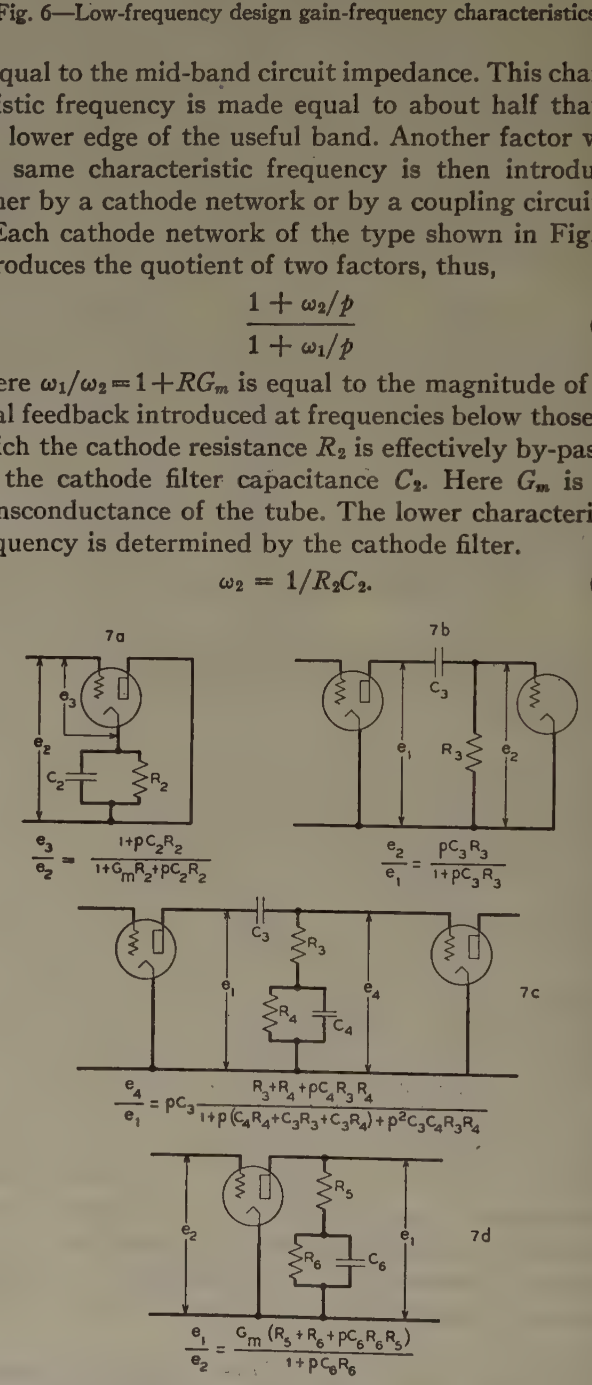

Fig. 7—Low-frequency circuits.
a—Cathode network
b and c—Coupling networks
d—Plate network

Each coupling circuit of the type shown in Fig. 7b, which consists of the coupling capacitance C_3 between the plate and the following grid and the grid resistance R_3, introduces a factor

$$\frac{1}{1+\omega_3/p} \tag{14}$$

where $\omega_3=1/R_3C_3$.

If the resistance R_3, determined by this equation, is appreciably less than the maximum allowable resistance on account of grid current, two additional factors may be introduced by adding a parallel resistance-capacitance circuit in series with the above grid resistance as shown in Fig. 7c.

This also changes the value of the first characteristic frequency so that the grid resistance must be adjusted to bring this characteristic frequency back to its desired value. The design, however, is straightforward. Given the coupling capacitance C_3 and the total grid resistance R_3+R_4, determined by other considerations, the highest critical frequency ω_3 for the network and the relation $\omega_3/\omega_4=\omega_4/\omega_5$ required to give the desired slope, then

$$R_3 = \frac{R_3+R_4}{\omega_3C_3(R_3+R_4)-1+1/[\omega_3C_3(R_3+R_4)]} \tag{15}$$

$$C_4 = C_3(R_3+R_4)^2/R_3R_4 \tag{16}$$

and the network introduces the factors

$$\frac{(1+\omega_4/p)}{(1+\omega_3/p)(1+\omega_5/p)} \tag{17}$$

where

$$\omega_3/\omega_4 = \omega_4/\omega_5 = \omega_3C_3(R_3+R_4). \tag{18}$$

Each plate circuit of the type shown in Fig. 7d introduces the quotient of two factors in such a way that the low-frequency gain is greater than the mid-band gain by the ratio of the total resistance in the plate circuit to the mid-band impedance. The lower characteristic frequency is determined by the time constant of the filter (R_6C_6) and the higher characteristic frequency is determined by the capacitance of the filter and the parallel combination of the filter resistance and the mid-band resistance. This assumes that the impedance of the coupling network is large compared with the remainder of the plate circuit as is usually the case in practice.

When these characteristic frequencies are properly located the desired over-all frequency characteristic may be obtained.

II. Feedback Demodulator

As has been remarked previously, the demodulator is in the β circuit so that any distortion produced in it will be present in the transmitter output. See Fig. 1. In order to minimize the distortion, a high-level linear rectifier may be used. This should have a large filament emission with small transit time and work into a high-impedance load. These requirements are not compatible with those for the maximum amount of feedback, which are low tube capacitance and a low output impedance. In order to satisfy these conflicting requirements J. W. McRae devised a circuit which effectively applies local feedback to the linear rectifier as shown in Fig. 1. This circuit allows the use of a small rectifier tube with small transit time, low capacitance, and only moderate filament emission working into a low output impedance. The effect of the feedback is to make the rectifier operate as if it had a high output impedance so that large filament emission is not required for small distortion. Though without feedback a very high physical resistance might be used, it would so increase the time constant that nontracking in the diode might result. The feedback reduces the time constant by a large factor. Demodulators have been built according to this circuit that operated successfully with 30 and 40 decibels of feedback.

It was found necessary to have a rather imposing array of radio-frequency circuits associated with the main β circuit. Circuit A of Fig. 1 acts as a shunt to the signal frequency that is present along with the modulated radio frequency on the modulator plates. Its presence is necessary in order that the output be fed back by means of modulation and demodulation and not directly at the signal frequency. Circuit B prevents the radio frequency from getting on the amplifier plate and helps prevent it from getting through to the grid of the first tube of the signal frequency amplifier. Choke C and series-tuned circuit D act as a filter for the amplifier grid. Finally, choke E suppresses the remaining radio-frequency voltage to a value which will not cause distortion in the first signal-frequency amplifier grid.

Receiver

The design and construction of the multiplex receiver are described in a companion paper.[12] It is a triple detection set. The first demodulator converts the radio-frequency carrier from 160 megacycles per second to about 8 megacycles per second at which intermediate frequency there is one stage of amplification. This is followed by the second demodulator that converts the carrier to a frequency at which it is convenient to build networks to discriminate against undesired signals such as the adjacent radio channel. The third demodulator recovers the type-K signal in the frequency range from 12 to 60 kilocycles per second.

In order to minimize the distortion the final demodulator is made a high-level linear rectifier. Undistorted power to drive this demodulator is obtained by employing negative feedback in the second intermediate-frequency amplifier. The proper level to the final demodulator is maintained by employing automatic volume control.

One of the major requirements of the receiver is that its phase characteristic be symmetrical with respect to the carrier. This requires that the interstage circuits be accurately tuned and that the frequencies of the beating

[12] D. M. Black, G. Rodwin and W. T. Wintringham, "Ultra-short-wave receiver for the Cape Charles-Norfolk multiplex radio-telephone circuit," Proc. I.R.E., this issue, pp. 95–100.

oscillators be accurately maintained. Toward this end both beating-oscillator frequencies are obtained from a crystal-controlled oscillator and harmonic-generator chain, in fact, from the same crystal oscillator. Even by employing crystal-controlled oscillators at both the transmitter and the receiver it was not deemed feasible in this particular apparatus to maintain the second intermediate-frequency carrier with sufficient accuracy to maintain the desired symmetry without taking steps to have the frequency characteristic linear.

DISTORTION MEASUREMENTS

Nonlinear Distortion

Fig. 8 shows the second-harmonic distortion from the transmitter *A* without feedback, *B* with feedback by

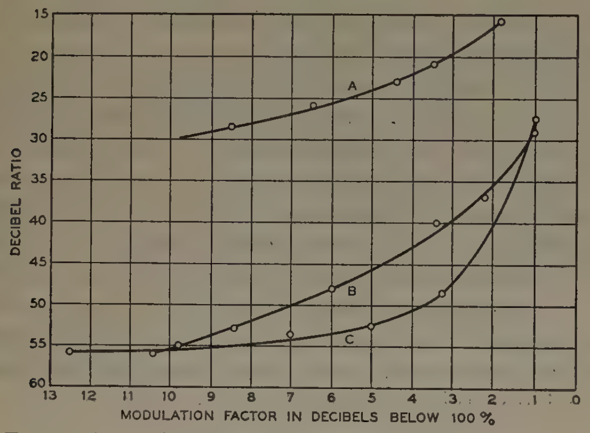

FIG. 8—Second-harmonic distortion from the transmitter.
A—Without feedback
B—With simple diode demodulation
C—With McRae's demodulator

linear diodes, and *C* with feedback using McRae's feedback demodulator. Similar tests were made for third-harmonic distortion and second- and third-order distortion with two input tones. These indicate that the crosstalk would be prohibitive without either feedback or the use of a spread-band system so that the second-order distortion falls in a vacant frequency range.

Phase Modulation

In a plate-modulated transmitter, phase modulation may be generated by an impedance common to the input and bridge-neutralizing circuits or by variation in the electron transit time with modulation voltage. Phase modulation from either of these causes was reduced by a redesign of the 356A tube. Phase modulation that was produced by the latter was reduced by diminishing the transit time, and that by the former was lessened by reducing the impedance common to the neutralizing circuit and input or output circuit. Besides reducing the length of leads within the tube, separate grid and plate leads were provided for the neutralizing circuit. Fig. 9 shows a picture of the new 364A tube having these desirable characteristics together with its predecessor, the 356A.

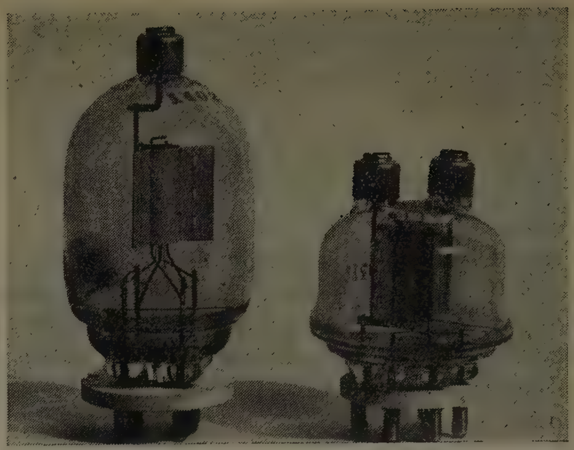

Fig. 9—Photograph of tubes used in first and final experimental models of the modulator.

Fig. 10 shows the phase modulation produced by the modulator with different tubes and circuit connections. The abscissas give the amplitude-modulation factor in decibels below unity. The ordinates give the phase-modulation index in degrees. The top curve shows the phase modulation produced by the 356A's and the bottom curve shows the phase modulation by the 364A's. The intermediate curves are the result of an attempt to evaluate the relative importance of the various factors involved. Curve *B* shows the improvement of the smaller grid-filament spacing and shorter leads without taking advantage of separate grid and plate leads. The other curves show the effect of using either separate grid or separate plate leads. While there is an unmistakable improvement in the new tube, it is difficult to say what part of it is the result of any particular change. This is partially due to the fact that the phase modulation depends on both the neutralizing condenser

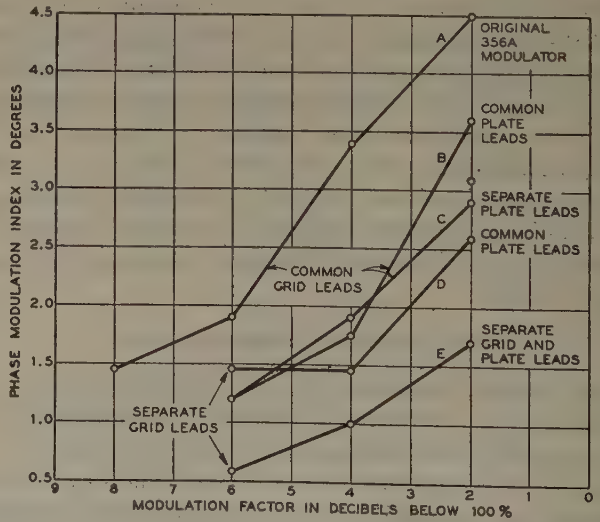

Fig. 10—Phase modulation as a function of amplitude-modulation factor.
A—Original 356A modulator
B—364A with common grid and plate connections.
C—364A with common grid and separate plate connections.
D—364A with separate grid and common plate connections.
E—New modulator using separate grid and plate connections on 364A.

etting and tuning of plate and grid circuits. In this espect the new tube is better than the original one.

The distortion produced by departures of the transmission characteristic of the receiver from the ideal was determined by measuring the over-all distortion of the system.

System Tests

As a practical test on the radio system, the experimental model was tested in the laboratory in connection with the type-K carrier equipment. The equipment arrangement for these tests is shown in Fig. 11. Type-K transmitting terminal with means for applying speech

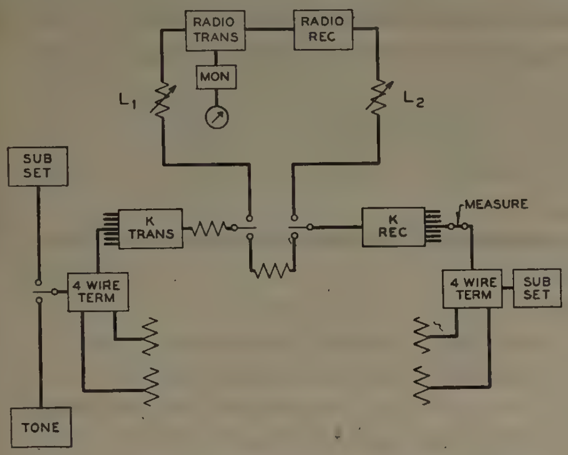

Fig. 11—Equipment arrangement for system tests.

or a single frequency is at one end, and the type-K receiving terminal with means for listening or measuring is at the other end. These two ends may be connected by means of an attenuator or through the radio repeater. By adjusting the attenuators L_1 and L_2 the level at which the radio equipment is operated can be controlled. A monitor is provided for the transmitter which allows the determination of the level at which the radio equipment is being operated. A significant way of expressing this level is to state the modulation factor for a single-frequency tone equal to the required load capacity at any previous point in the circuit. This modulation factor is called B when it is expressed in decibels below unity. Increasing attenuator L_1 decreases the modulation factor on the transmitter and hence decreases the distortion. In order to maintain the same over-all net loss, however, it is necessary to decrease attenuator L_2 by an equal amount. This increases the first circuit noise from the receiver. Hence, it is desirable to operate the radio equipment at as high a level as possible without exceeding the distortion requirements.

Intermodulation tests were made both with test tones and with speech. In each case the signal was applied to one or more channels and the intermodulation measured in the remaining channels. Measurements were made with the radio equipment operated at various levels. The results of the more significant intermodulation with the first experimental model of the radio equipment operating at the level $B=5$ are given in Figs. 12 and 13. Curve A represents the design limit below which it was desired to keep the distortion in the radio transmitter for this particular installation. The curves of

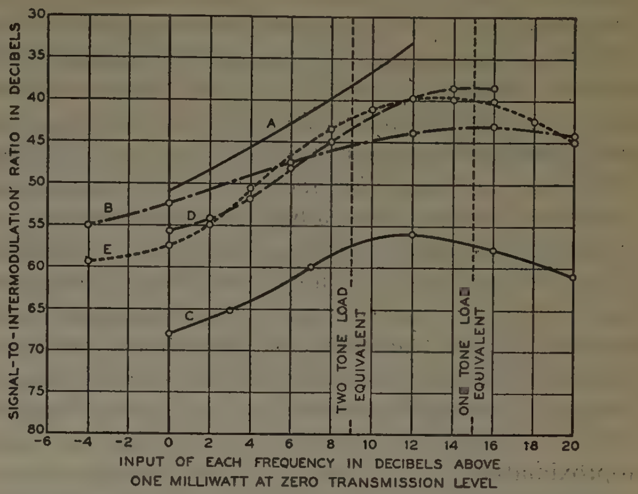

Fig. 12—Distortion characteristics of the first model. The abscissa gives the input power of each test tone in decibels above 1 milliwatt at the transmitting toll testboard. The radio equipment for these tests was lined up so that a single tone of +20 decibels referred to a milliwatt or two tones each of +14 decibels would modulate the transmitter 100 per cent. The peak load requirement for a 12-channel telephone system is +15 for a single tone or +9 for each of two tones. The ordinate gives the ratio of the distortion to each test tone in decibels below unity so that a rising curve indicates increasing distortion.

A—Intermodulation design limit.
B—Second-harmonic distortion at 57.0 kilocycles per second in channel 12.
C—Second-harmonic distortion at 25.0 kilocycles per second in channel 4.
D—Second-order cross modulation at 13 kilocycles per second (=58−45).
E—Third-order cross modulation at 49 kilocycles (=2×53−57).

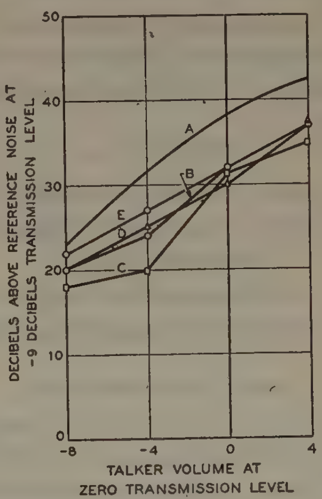

Fig. 13—Talking Tests as measured with 2*B* noise meter.
A—Cross modulation design limit.
B—Cross modulation in channel 12 from channel 5.
C—Cross modulation in channel 10 from channels 11 and 12.
D—Cross modulation in channel 11 from channels 4 and 5.
E—Cross modulation in channel 1 from channels 9 and 12.

Figs. 12 and 13 indicate that the experimental radio equipment meets this requirement. The performance illustrated by the curves is obtained when the transmitter is operated at a level such that a sine wave equal

to the peak load of 12 telephone channels will modulate the transmitter 5 decibels below complete modulation. As a matter of interest, the single-frequency and two-frequency inputs equivalent to the 12-channel load are indicated in Fig. 12. The predominant intermodulation involves a high-frequency channel. On comparing, for example, curves B and C, it is seen that the second-harmonic distortion occurring at 57 kilocycles per second is very much greater than that occurring at 25 kilocycles per second. The fact that the predominant intermodulation involves a high-frequency channel was interpreted as being caused by either phase modulation in the transmitter, or a nonsymmetrical transmission characteristic in the receiver, or both. Accordingly, the modulator of the transmitter was redesigned and the intermediate-frequency circuits of the receiver were redesigned. These changes reduced the second-harmonic intermodulation at 57 kilocycles per second to a value approximately equal to that at 25 kilocycles per second; that is, curve C of Fig. 12 represents the second-harmonic intermodulation on any test tone after the changes were made. Similar tests on second- and third-order intermodulation products indicate that the modified transmitter may be operated at a 3-decibel higher level without increasing the intermodulation over that of the first experimental model.

Besides the reduction in distortion and phase modulation with the new transmitter, there is an increase in carrier output from 30 to 80 watts so that the undistorted sideband power is 7 decibels more than the first model.

ACKNOWLEDGMENT

We wish to express our appreciation of the support and encouragement given in the course of this work by Mr. J. C. Schelleng, under whose direct supervision the work was done.

APPENDIX

Cross Modulation Resulting from Deviations of the Gain-Phase-Frequency Characteristic from the Ideal

Consider an ideal transmitter and an ideal linear rectifier but a receiver whose gain and phase is proportional to

$$A_x \exp i(k_1\omega + k_2 x + \phi_x) \tag{19}$$

at any frequency $\omega + x$. Here ω is the carrier frequency and k_1 and k_2 are constants, and ϕ_x is the departure of the phase from linearity. Here the implied factor of proportionality is adjusted so that $A_0 = 1$ and k_1 is chosen so that $\phi_0 = 0$. Let the transmitter output be

$$\left\{1 + m_p \frac{\epsilon^{ipt} + \epsilon^{-ipt}}{2} + m_q \frac{\epsilon^{iqt} + \epsilon^{-iqt}}{2}\right\} \epsilon^{i\omega t} \tag{20}$$

representing a carrier of constant frequency (no phase modulation) ω, whose envelope perfectly reproduces the input signal,

$$m_p \epsilon^{ipt} + m_q \epsilon^{iqt}. \tag{21}$$

The input to the linear rectifier is then proportional to

$$\left\{1 + \frac{m_p}{2}\left(A_p \epsilon^{ip(t+k_2)+i\phi_p} + A_{-p}\epsilon^{-ip(t+k_2)-i\phi_{-p}}\right) + \frac{m_q}{2}\left(A_q \epsilon^{iq(t+k_2)+i\phi_q} + A_{-q}\epsilon^{-iq(t+k_2)-i\phi_{-q}}\right)\right\} \epsilon^{i\omega(t+k_1)}. \tag{22}$$

This may be written in the form

$$(1 + a + ib)\epsilon^{i\omega(t+k_1)}. \tag{23}$$

The output of the linear rectifier is then proportional to the magnitude of the coefficient of $\exp i\omega(t+k_1)$.

$$\begin{aligned} |1 + a + ib| &= (1+a)\sqrt{1 + b^2/(1+a)^2} \\ &= 1 + a + (b^2/2)(1 - a + a^2 - a^3 + \cdots) \\ &\quad - (b^4/8)(1 - 3a + 6a^2 - 10a^3 + \cdots) \\ &\quad + \cdots. \end{aligned} \tag{24}$$

The predominant fundamental and second order terms are

$$\begin{aligned} &(m_p/2)(A_p\epsilon^{i\phi_p} + A_{-p}\epsilon^{-i\phi_{-p}})\epsilon^{ipt_2} \\ &+ (m_q/2)(A_q\epsilon^{i\phi_q} + A_{-q}\epsilon^{-i\phi_{-q}})\epsilon^{iqt_2} \\ &+ (m_p m_q/8)(A_p\epsilon^{i\phi_p} - A_{-p}\epsilon^{-i\phi_{-p}})(A_q\epsilon^{i\phi_q} \\ &\qquad - A_{-q}\epsilon^{-i\phi_{-q}})(\epsilon^{i(p+q)t_2} + \epsilon^{i(p-q)t_2}) \\ &+ (m_p^2/16)(A_p\epsilon^{i\phi_p} - A_{-p}\epsilon^{-i\phi_{-p}})^2\epsilon^{i2pt_2} \\ &+ (m_q^2/16)(A_q\epsilon^{i\phi_q} - A_{-q}\epsilon^{-i\phi_{-q}})^2\epsilon^{i2qt_2}. \end{aligned} \tag{25}$$

Here $t_2 = t + k_2$, where k_2 is the envelope delay.

This shows that in order to prevent distortion (intermodulation) in a double-sideband system the gain-phase-frequency characteristic must be symmetrical about the carrier frequency. Let

$$\begin{aligned} \theta_p &= \phi_p + \phi_{-p} \qquad & \delta_p &= A_p - A_{-p} \\ \theta_q &= \phi_q + \phi_{-q} \qquad & \delta_q &= A_q - A_{-q} \end{aligned} \tag{26}$$

represent the deviations of the gain-phase-frequency characteristic from symmetry. Here θ_p and θ_q represent the difference between the phase at the upper side frequency from the straight line drawn through the phase at the lower side frequency and the phase at the carrier. δ_p and δ_q represent the difference between the gains expressed as ratios at upper and lower side frequencies. If these deviations are small, the output of the system will be

$$\begin{aligned} &m_p\epsilon^{ipt} + m_q\epsilon^{iqt} \\ &\quad + (m_p m_q/8)\sqrt{(\theta_p^2 + \delta_p^2)(\theta_q^2 + \delta_q^2)}\,(\epsilon^{i(p+q)t} + \epsilon^{i(p-q)t}) \\ &\quad + (m_p^2/16)(\theta_p^2 + \delta_p^2)\epsilon^{i2pt} + (m_q^2/16)(\theta_q^2 + \delta_q^2)\epsilon^{i2qt}. \end{aligned} \tag{27}$$

Editor's Note: This paper was prepared in March, 1941, but it was not submitted for publication until August, 1944, because of security reasons.

Ultra-Short-Wave Receiver for the Cape Charles-Norfolk Multiplex Radiotelephone Circuit*

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Summary—**The requirements for an ultra-short-wave receiver for use in a multiplex radiotelephone link circuit are outlined. The technical details of a receiver designed to meet such requirements in the circuit between Cape Charles and Norfolk, Virginia, are described.**

Introduction

THE development of the receiver described in this paper was undertaken as part of a program to provide practical receiving equipment for use in unattended multiplex radiotelephone links in toll telephone systems. Single-channel ultra-short-wave radio links over short distances have been in operation for a number of years;[1–3] multichannel systems for similar types of service have also been in use.[4,5] Previous multiplex radio systems had fewer channels than this one and as a general rule the channels were spread about the carrier frequency in such a way as to give a minimum amount of crosstalk. The system incorporating the receiver described in this paper differs from these in that there are 12 channels placed in a solid compact group appearing on both sides of the radio carrier in a double-sideband type of transmission. The radio equipment is designed to operate between standard type-K wire-carrier telephone terminals, in which the 12 individual single-sideband channels at 4-kilocycle intervals are located in the frequency spectrum between 12 and 60 kilocycles.

Receivers built in accordance with the principles discussed in this paper have been in continuous service in the multiplex radiotelephone circuit between Cape Charles and Norfolk, Virginia,[6] which was opened to commercial traffic in October, 1941. In the radio circuit a transmitted power of 50 watts in the range of frequencies between 150 and 160 megacycles is used to obtain the desired signal strength at the receiving end 26 miles away. Directive antennas on 198-foot towers provide an optical path over the intervening Chesapeake Bay.

Circuit Description

The receiver circuit, which is of the triple-detection type, is shown in block schematic form in Fig. 1. The signal is supplied directly to the first detector from the antenna system, and is there converted to the first intermediate frequency of approximately 10,000 kilocycles. After one stage of amplification it is then converted to 1500 kilocycles in the second detector. The heterodyne frequencies applied to the first and second detectors are obtained from a single temperature-controlled crystal oscillator through a series of harmonic generators and amplifiers. The output of the loop-test oscillator can be supplied to the first detector for making local tests when the distant transmitter is not operating.

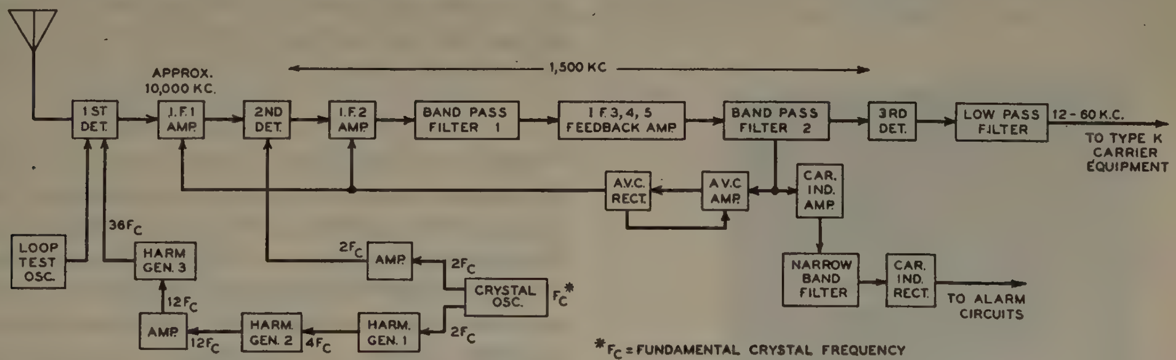


Fig. 1—Block schematic multiplex receiver.

At the 1500-kilocycle frequency another single-stage amplifier, like the one at the higher intermediate frequency, has its gain varied by the automatic-volume-control circuit. In addition, there is a three-stage fixed-

* Decimal classification: R361. Original manuscript received by the Institute, August 7, 1944. Presented, Winter Technical Meeting, New York, N. Y., January 27, 1945.
† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] A. G. Clavier, "Production and utilization of micro-rays," *Elec. Commun.*, vol. 12, 1, p. 3; July, 1933.

[2] A. J. Gill, "Interim report on ultra-short-wave experimental transmissions between Guernsey and the mainland," Post Office Engineering Department, Radio Report No. 269, July 21, 1934.

[3] N. F. Schlaack and F. A. Polkinghorn, "An unattended ultra-short-wave radiotelephone system," Proc. I.R.E., vol. 23, pp. 1275–1285; November, 1935.

[4] D. B. Mirk, "Belfast-Stranraer 9-circuit ultra-short-wave radiotelephone system—Part II," *Post Office Elec. Eng. Jour.*, pp. 33–40; April, 1938.

[5] S. Matsumae and S. Yonezawa, "Equipments for multiplex carrier telephony on ultra-short wave," *Nippon Elec. Commun. Eng.*, no. 15, pp. 554–563; February, 1939.

[6] N. F. Schlaack and A. C. Dickieson, "Cape Charles-Norfolk ultra-short-wave multiplex system," Proc. I.R.E., this issue, pp. 78–83.

gain feedback amplifier and two sections of band-pass filters to provide the additional amplification and selectivity required. The automatic-volume-control circuit consists of an amplifier and rectifier. The rectifier feeds back negative voltage in the usual manner to the two stages in the receiver. This same rectifier also supplies a voltage of a positive polarity to its own amplifier. This gives a very flat volume-control characteristic. Another branch circuit has an amplifier, narrow-band filter, and rectifier, together with an alarm relay to indicate carrier failure or frequency drift. The third detector is a high-amplitude diode working into a low-pass filter whose output is connected directly to the type-K carrier telephone equipment.

Vacuum-tube currents in all tubes in the receiver, with the exception of those in the power supplies, can be measured by a meter located on one of the panels. A multiposition switch permits the meter to be connected to any one of forty separate circuits. The meter measures the voltage drop across a resistance in the plate, screen, or cathode lead so that the circuits are not opened at any time.

Three separate power supplies are provided in the receiver, all of which are of the regulated type which give constant output voltage with varying load and line-voltage conditions. A 180-volt power supply with the positive side grounded is used to provide the negative bias required for the harmonic generators and other tubes. A relay connected in the output of this power supply prevents either of the other power supplies from operating until the full negative voltage is built up.

Another 180-volt power supply is used to provide the plate and screen voltages for most of the other tubes in the receiver. A third power supply is used to provide the 400 volts required for the crystal-oscillator tubes and the feedback amplifier. A regulating transformer is used to provide constant filament voltage on all tubes in the receiver proper for a range of line voltage from 90 to 130 volts. An external source of 24 volts direct current is required to operate the alarm and testing relays in the receiver.

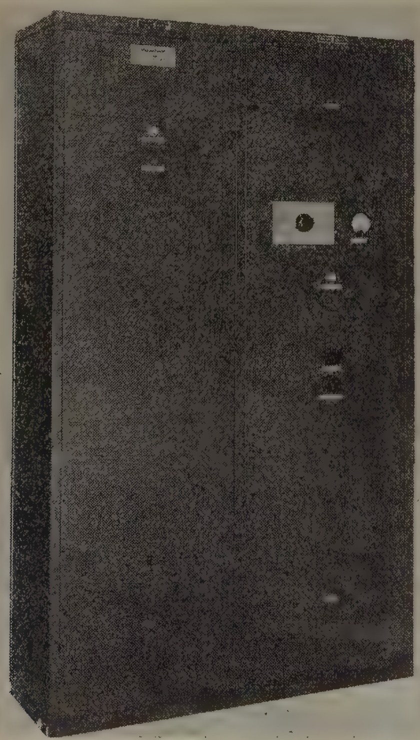


Fig. 2—Front view of receiver.

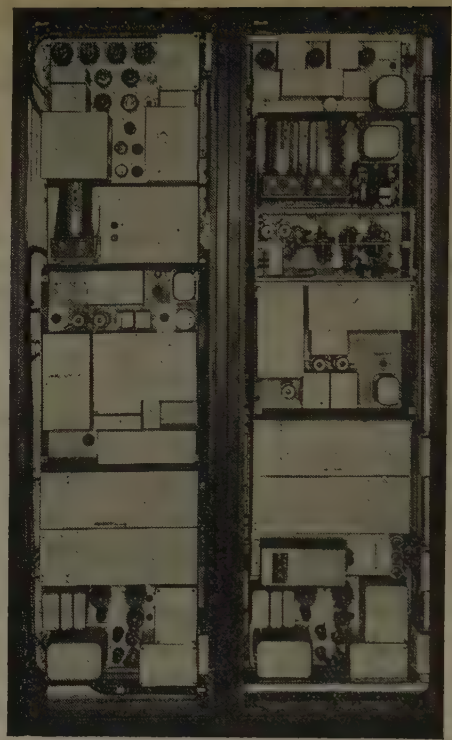


Fig. 3—Rear view of receiver.

Mechanical Description

The mechanical construction of the receiver follows the usual practice of mounting individual panels in enclosed metal cabinets. Two views of the receiver are illustrated in Figs. 2 and 3. The receiver consists of two cabinets bolted together on a metal base. The panels are of the depressed type with apparatus mounted on the back side and the terminals and wiring in the space between the front of the panel and the mat. The mats which cover the front of the panels present a uniform flat surface matching the other part of the cabinets which have an aluminum-gray finish.

Since the receiver is designed for a fixed frequency, no tuning dials are provided on the front of the panels. Such tuning adjustments as are necessary can be made with a screwdriver inserted in the appropriate hole, which is normally covered with a small plug, in the front mat. The tuning controls have locking devices to prevent them from getting out of adjustment. The individual circuits are enclosed in shield cans and the separate panels in most cases have large cans enclosing all of the apparatus mounted on the back side. The covers of the cans as well as the front mats are equipped

with safety switches so that when they are taken off, the high voltages are removed from any exposed terminals. All alternating and high-voltage wires between panels are contained in armored cable or pipe conduit.

Requirements and Design Considerations for a Multiplex Receiver

The reasons for the choice of the double-sideband method of transmission for this system are contained in a companion paper.[7] As with any other telephone system, the performance standards of any part of the system, such as the receiver, are so proportioned that each part carries an economical share of the burden. The performance standards to which this receiver was designed are discussed in the following paragraphs.

From the point of view of the designer of the multiplex radiotelephone system, the receiver is a box with input terminals to which an antenna is connected and output terminals to be connected to a telephone line. The performance of the receiver is of interest to him only as it reacts on the system as a whole. It is the job of the designer of the receiver to fill the box in such a way that the performance measured between its terminals is satisfactory. The general requirements for the multiplex receiver, therefore, only describe the performance desired between its input and output terminals.

In this system, the receiver is required to accept an amplitude-modulated double-sideband signal from the antenna transmission line and deliver the signal consisting of 12 single-sideband channels at frequencies from 12 to 60 kilocycles to a telephone cable pair. It was estimated that the carrier voltage across a 72-ohm coaxial line from the antenna, when properly terminated, would be about 2500 microvolts. Some variation from this nominal voltage would be expected due to aging of the transmitter tubes, etc. Under full-load conditions with all 12 channels in service the peak-modulation index of this carrier would approach 50 per cent.

The receiver should deliver an output to the carrier system about 40 decibels below 1 milliwatt for each individual channel. There is a further requirement that the output of the receiver for an individual channel should not vary more than ± 0.5 decibel. To compensate for all the gain variations between the input to the transmitter and the output of the receiver, automatic volume control is required in the receiver. In order to keep the loss of individual channels to a constant value, the gain of that portion of the receiver preceding the final detector must be flat over a band somewhat greater than 120 kilocycles.

In any radio system nonlinear distortion is present to a greater or lesser degree. In a single-channel system, intermodulation produced by such nonlinearity is of importance only to the extent that distortion of the wave forms of the signals is produced and impairs the quality of speech. In a multichannel system, however, nonlinear distortion produces intermodulation products which can appear in channels other than those in which the fundamental signals are transmitted, and hence cause crosstalk or noise. The requirements for the reduction of such crosstalk and noise are very much more severe than those for distortion within a single channel. The difference in severity of the requirements arises because in the single-channel case the intermodulation products can be present only when the signal is present and therefore tend to be masked by it. In a multichannel system, however, the intermodulation products may fall into a channel which is in use but at the moment is not carrying speech. In this case, the only masking element is the background noise in the channel.

In the present double-sideband system, the use of multiplex imposed requirements on nonlinear distortion many times more severe than is required for single-channel systems. Some sources of distortion not usually considered, became important in the receiver. The carrier voltage applied to the diode detector was made high to minimize the curvature of the diode characteristic and the impedance of the diode-load circuit was made a substantially constant resistance to all of the sideband frequencies further to reduce the distortion produced in the detector. The most difficult source of nonlinear distortion to control, however, was the shape of the amplitude and phase characteristics of the band-pass filters in the intermediate-frequency part of the receiver.

In examining the filter characteristics as a source of distortion, it is necessary to include the action of the final detector. It is well known that a perfect linear detector will reproduce the amplitude modulation of a carrier wave without distortion. However, the same linear detector is likewise known to generate distortion when it is called upon to produce the beat frequency between a carrier and a single-sideband signal. Consider first the effect of an ideal filter on an amplitude-modulated carrier. The amplitude of the two sidebands relative to that of the carrier may be altered by the same amount, in which case the only result is a change in the modulation index or degree of modulation; or the phase of one sideband may be linearly advanced and the other linearly retarded by the same amount relative to the carrier, in which case the signal as a whole is advanced or delayed. In either of these two cases the presence of the filter makes no contribution to the nonlinear distortion. However, in the case of a nonideal filter, if the change in amplitude of one sideband through the filter is different from that of the other, or if the phase advance of one sideband differs from the phase retardation of the other, intermodulation products will be produced. The result, under this condition, may be pictured by considering the wave at the output of the filter as being made up of two components, one formed by the carrier with symmetrical sidebands, and the other being a single component at the frequency of one

[7] C. R. Burrows and A. Decino, "Ultra-short-wave multiplex," Proc. I.R.E., this issue, pp. 84–94.

sideband and of such amplitude and phase as to represent the difference between the sideband transmitted through the filter of this frequency and the symmetrical sideband component of the same frequency. The symmetrical components form an undistorted amplitude-modulated wave, the modulation on which may be recovered without distortion. The excess or difference component, however, will be demodulated against the perfectly modulated carrier as a single-sideband signal, with all of the distortion resulting from such demodulation by a linear detector.

More exact analysis confirms the conclusion of this very rough descriptive analysis. In order that the nonlinear distortion be kept within satisfactory limits for this multiplex system, the phase shift through the filters must be linear to within a few degrees, and the amplitude distortion over the pass band of the filter must be less than a decibel.

An analysis of the selectivity required in the receiver indicated that it should be of the triple-detection variety in which the signal is converted in frequency twice before being detected by its own carrier. It was thought that the importance of interference-free operation warranted a very high over-all selectivity. In line with previous experience in similar circumstances a value of 90 decibels discrimination against adjacent-channel interference was thought to be desirable. The last intermediate frequency was, therefore, chosen as low as was consistent with easily attaining a symmetrical amplitude and phase characteristic and designing the necessary by-pass circuits for the final detector, which will be discussed later. The first intermediate frequency was chosen roughly as the geometric mean between the signal frequency and the last intermediate frequency.

Ordinarily a linear detector would consist of a diode operating into a high resistance. The output would be by-passed at the intermediate frequencies by a condenser and the detector would be operated at a high input amplitude. With this receiver the band to be transmitted is 120 kilocycles requiring that the by-passing be good at the final intermediate-carrier frequency plus and minus 60 kilocycles, and at the same time the output impedance of the detector be uniform and high at the frequencies from 12 to 60 kilocycles. It was found that by making the final intermediate frequency 1500 kilocycles, a reasonable design could be obtained for both the final intermediate-frequency selectivity and the detector output-circuit-impedance characteristics.

In order to obtain a large signal input to the final detector in the receiver, it was necessary to build an intermediate-frequency amplifier which had a high gain and which could handle a large signal voltage with extremely low distortion. The amplifier needed to have a wide flat band in order to satisfy all the requirements. This made it desirable to build an amplifier having a large amount of negative feedback.

It is not possible to construct a feedback amplifier which has a broad pass band and sharp cutoff at the edges.[8] For this reason it was necessary to construct a feedback amplifier having a broad band and to limit the band by two filters, one placed ahead of the amplifier and the other after the amplifier.

The concentration of a large amount of fixed gain in the feedback amplifier to obtain good linearity put a limitation on the automatic-volume-control circuit, since there were only two stages of amplification in the receiver which could be varied by the automatic volume control. To provide a sufficiently constant output a third amplifier stage was provided ahead of the volume-control rectifier but not in the main signal path through the receiver. A voltage which becomes more negative as the input carrier increases is supplied to the first two stages in the conventional manner. A voltage which becomes more positive as the signal increases is supplied to the third amplifier. The effect of increasing the gain of the third amplifier as the signal becomes stronger is to increase the negative voltage supplied to the other stages, and thus reduce the gain in the main signal path. By properly proportioning voltages the receiver output can be kept practically constant over a wide range of input.

The frequency stability of both transmitter oscillator and receiver-heterodyne oscillator must be such that the signal at the last intermediate frequency operates on a suitable linear portion of the filter characteristic. A fairly small deviation of either frequency would introduce an intolerable amount of distortion in the output. This requirement makes necessary the provision of considerably greater frequency stability in the transmitter than would be required for keeping on the assigned channel.

Since the receiver was to be used at a fixed frequency, it was desirable to provide a single crystal oscillator to supply both heterodyne-oscillator frequencies required. For the first beating oscillator it was necessary to multiply the fundamental frequency of the oscillator with several harmonic-generator stages, as suitable commercial crystals of the required stability were not obtainable for frequencies above a few thousand kilocycles. The second beating oscillator made use of the second harmonic of the crystal oscillator. The use of a common oscillator to supply both beating oscillator frequencies imposed a fixed relationship between the first intermediate frequency and the received frequency. This resulted in the receivers at opposite ends of the circuit having slightly different first intermediate frequencies, since the operating frequencies in the two directions are not the same.

Since the frequency stability is so important, some means of detecting any large deviation of the transmitted carrier from the center of the band of the receiver was required. Accordingly, a narrow-band crystal

[8] H. W. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, 3, pp. 421–454; July, 1940.

filter was designed to be placed in a branch circuit connected to the second intermediate-frequency amplifier just ahead of the final detector. A relay in a rectifier circuit in this branch gives an alarm in case the carrier frequency drifts outside the band of the filter.

From an operating standpoint it was necessary in case of circuit failure to determine quickly at which end of the circuit the trouble has occurred. A simple way of doing this was to introduce a suitable voltage into the first detector from another oscillator of such frequency that the beat frequency between it and that of the local transmitter was exactly equal to the frequency of the distant transmitter. Since this frequency had to be held within close limits in order to pass through the narrow-band filter, another temperature-controlled crystal oscillator was required for this loop-test oscillator. The antenna system was designed so that the pickup into the receiver from the local transmitter would be great enough so that the combination of it with the additional oscillator in the first detector of the receiver would produce a signal in the receiver at least as strong and preferably two or three times stronger than that from the distant transmitter. At the same time the pickup should not be great enough to produce cross modulation difficulties in the receiver.

Since the failure of only one tube in the multiplex receiver may interrupt service on twelve separate circuits, it is very important to make frequent checks on the condition of the various tubes. In order to avoid interrupting the circuits by removing the tubes from the receiver to check them with a vacuum-tube tester, a scheme was developed to be able to check the tubes under operating conditions. A switch on one of the panels reduces the filament voltage on the tubes by 10 per cent. The currents measured at the metering panel under this condition indicate whether any of the tubes have low emission.

Performance

The curve shown in Fig. 4 illustrates the amplitude-frequency characteristic of the pass band of the two intermediate-frequency filters connected in tandem. The over-all frequency characteristic of the pass band of the receiver is essentially the same as the curve shown. This characteristic proved to be satisfactory in service.

In Fig. 5 the over-all frequency characteristic is plotted on a different scale to show the selectivity obtained outside of the nominal band of the receiver. The discrimination of about 90 decibels on adjacent channels has proved to be ample insurance against interference from signals near the operating frequency.

The curve in Fig. 6 shows only the departure from linearity which was obtained in the over-all phase characteristic of the receiver, since this is the criterion which determines the contribution to nonlinear distortion. If a curve of total phase shift were shown instead, the departure from linearity would not be evident, since

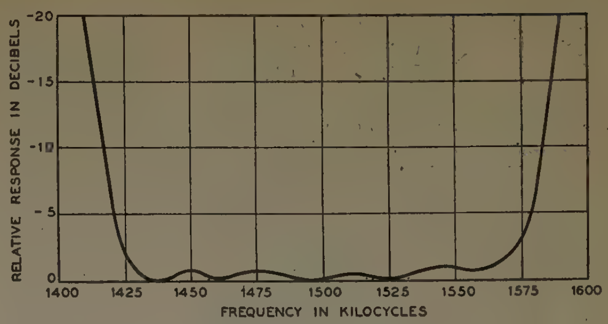


Fig. 4—Amplitude-frequency characteristic pass band of two intermediate-frequency filters in tandem.

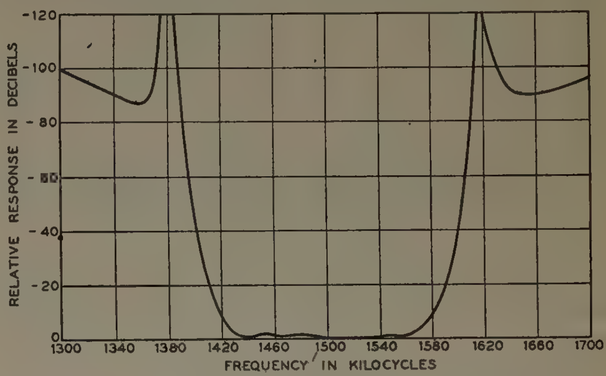


Fig. 5—Over-all amplitude-frequency characteristic.

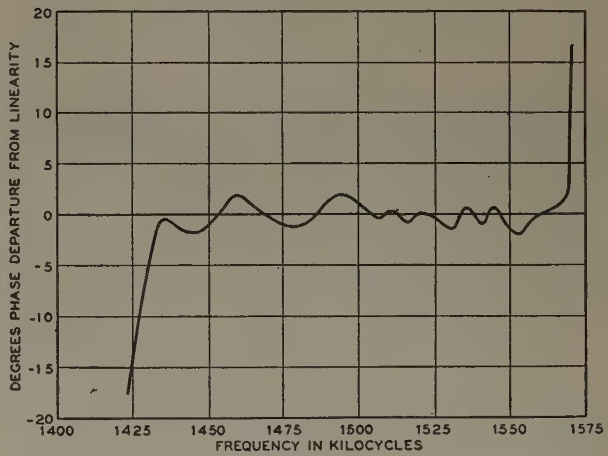


Fig. 6—Over-all phase-frequency characteristic.

the phase shift in the filters alone amounts to about 180 degrees in the middle of the band and has a slope of about 16 degrees per kilocycle. About two thirds of the amount of phase departure from linearity is contributed by the two filters and one third contributed by the other tuned circuits in the receiver. This departure is within the value which was being striven for, and it was found that it was not possible to improve it materially without increasing the complexity of the filter design beyond practical limits.

The low-pass filter in the output of the final detector has an amplitude characteristic as shown in Fig. 7. The group of 12 channels from 12 to 60 kilocycles pass through this filter and are supplied to the type K carrier equipment at substantially equal volumes.

The automatic-volume-control characteristic is shown in Fig. 8. The use of positive-voltage feedback on the

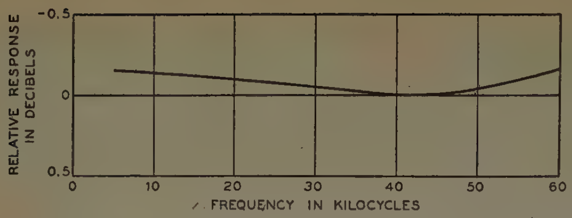


Fig. 7—Amplitude-frequency characteristic third-detector output filter.

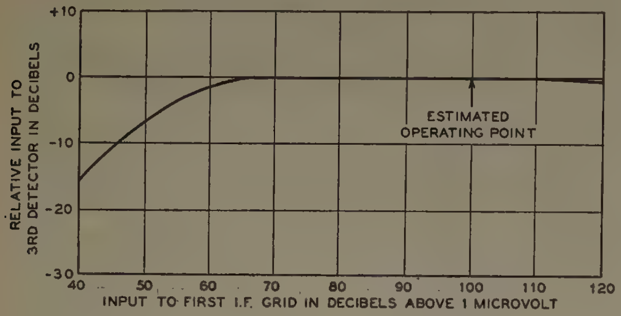


Fig. 8—Automatic-volume-control characteristic.

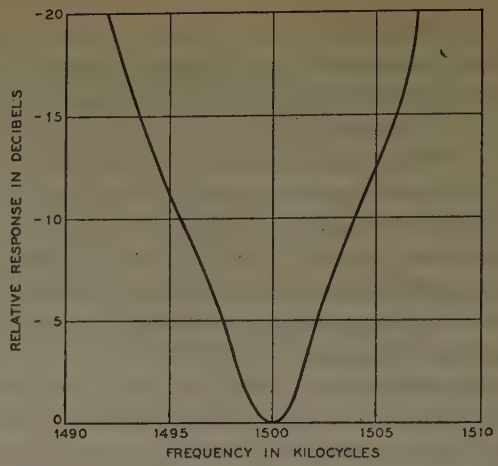


Fig. 9—Amplitude-frequency characteristic carrier-indicator filte

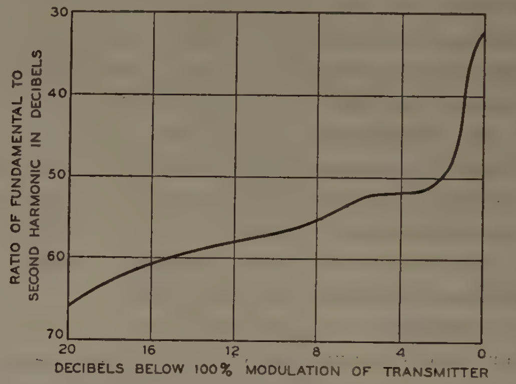


Fig. 10—Typical over-all signal-to-distortion ratio 28.5 kilocycles modulating frequency.

automatic-volume-control amplifier made possible the flat characteristic over such a wide range of input with control on only two stages in the intermediate-frequency amplifier. With the operating point as shown, the incoming signal can drop more than 30 decibels before the receiver-output volume will change appreciably.

The narrow-band crystal filter used to indicate departures of the carrier from its correct frequency has a characteristic illustrated in Fig. 9. The carrier-indicator relay is set to give an alarm when the amplitude drops 6 decibels. This will occur if the frequency of the carrier is off by as much as 2.5 kilocycles.

The performance of the receiver in regard to noise or distortion is difficult to measure independently of the other parts of the radio circuit. As an example of the results which were obtained in regard to distortion, a typical curve including the transmitter and measured at the output of the receiver is illustrated in Fig. 10. Signal-to-distortion ratio is plotted against percentage modulation in the transmitter. In actual operation o the radio circuit an input of 1 milliwatt at the tol switchboard on one of the twelve channels modulates the transmitter about 20 decibels below 100 per cent modulation so that it is very seldom that the peak volume of all twelve channels approaches full modulation. The signal-to-noise ratio of the over-all circuit including the transmitter and associated carrier telephone equipment measures approximately 60 decibels in each of the single channels at normal volume.

Correction

Dr. William B. Rogers has drawn the attention of the editors to the following corrections to his paper, "Electronic Apparatus for Recording and Measuring Electrical Potentials in Nerve and Muscle," which appeared on pages 738 to 743 of the December, 1944, issue of the PROCEEDINGS: The second paragraph in the right-hand column on page 738 should read "Directly coupled to the oscillator is a 6J7 tube (T_2) which amplifies the *pulse* resulting from the *charging current*."

The fifth paragraph in the right-hand column of page 738, fifth line, should read "The *charging pulse* amplified by a 6J7 tube (T_1), . . ."

Ultra-Short-Wave Transmitter for the Cape Charles-Norfolk Multiplex System*

R. J. KIRCHER†, SENIOR MEMBER, I.R.E., AND R. W. FRIIS†, SENIOR MEMBER, I.R.E.

Summary—**Design features of an unattended ultra-short-wave double-sideband multiplex transmitter are described. Forty decibels of envelope feedback is utilized over the 12- to 60-kilocycle band of the twelve type-K carrier-signal channels which modulate the last stage of the transmitter. Accessibility of apparatus and ease in maintenance contribute toward obtaining maximum reliability of the equipment in commercial service.**

RADIO equipment in the 160-megacycle region has been developed as a means for spanning natural barriers which present serious obstacles to the normal cable and wire routes. For this service a number of severe requirements are imposed on the radio apparatus, particularly when it is to operate on an unattended basis. For example, an absolute frequency stability of approximately 0.002 per cent is required in the radio link. A further requirement is that nonlinear distortion products which appear as crosstalk should be suppressed approximately 50 decibels below the fundamental frequencies from which they are derived. From the operational standpoint, service interruptions must be guarded against in every way possible, and the utmost convenience in servicing and maintaining the equipment provided. This paper describes the transmitter developed for this purpose which meets these requirements.

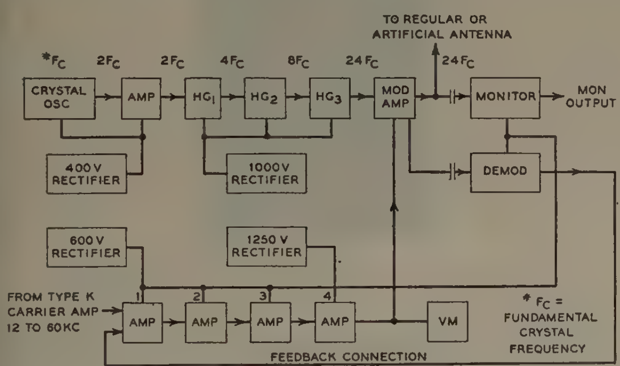


Fig. 1—Block schematic of the transmitter.

The most significant feature in the design of the transmitter is the use of 40 decibels of envelope negative feedback which is effective in the 12- to 60-kilocycle transmission band. The Chesapeake Bay installation of this equipment is discussed in a companion paper.[1]

By referring to the block schematic diagram of the radio transmitter given in Fig. 1, a general idea of the components may be obtained. The radio-frequency carrier is produced by a chain of units consisting of a quartz-crystal oscillator, a buffer amplifier, and three stages of harmonic generators. The last stage, which is the modulator stage, is a bridge capacitance neutralized, plate-modulated, class C amplifier which delivers approximately 50 watts of radio-frequency power. This stage may be modulated 100 per cent. The signal-modulating power is developed through a chain of four class A amplifier stages. Loosely coupled to the modulator is the demodulator unit from which the signal envelope voltage of correct phase is obtained which is returned to the input of the signal amplifying stages to effect the negative-feedback correction on the transmitter. A monitor unit designed to check the operation of the transmitter is connected at the junction of the output circuit of the modulator and the balanced coaxial transmission line. With this general picture as a guide, a more detailed description of the equipment and its operation follows.

Mechanical Design

Maximum accessibility and ease in maintenance without any sacrifice in operating performance must be

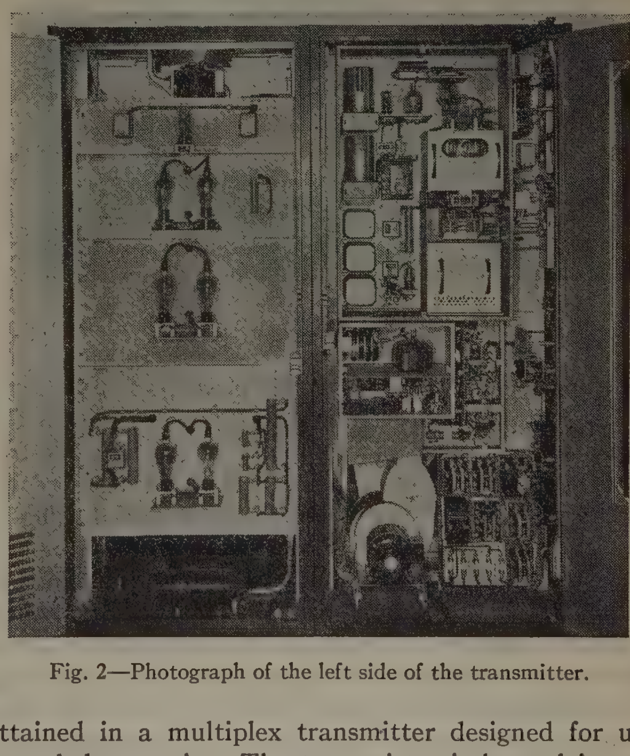


Fig. 2—Photograph of the left side of the transmitter.

attained in a multiplex transmitter designed for unattended operation. The transmitter is housed in two 6-foot steel cabinets each 30 inches deep and $21\frac{1}{2}$ inches wide which are bolted together. Each cabinet has a full-length door on the right and left sides; the left-side door in each case has a long glass window through which most of the tubes within may be seen. The inside of the cabinets, and most of the apparatus, are finished with

* Decimal classification: R355.5. Original manuscript received by the Institute, August 7, 1944. Presented, Winter Technical Meeting, New York, N. Y., January 27, 1945.
† Bell Telephone Laboratories, Inc., New York, N. Y.
[1] A. C. Dickieson, N. F. Schlaack, "Cape Charles-Norfolk ultra-short-wave multiplex system," Proc. I.R.E., this issue, pp. 78–83.

aluminum lacquer to provide good interior illumination. Most of the apparatus is mounted on both sides of flat 27-inch panels supported by relay-rack channels centrally placed on the front and back of the cabinets. The wide panels allowed a greater flexibility in the initial design, and with an equipment depth of approximately 11 inches, a high degree of accessibility is realized.

A full left-side exposure of the transmitter appears in Fig. 2. The cabinet on the left contains the four rectifiers and the regulating transformers for the 115-volt alternating-current primary power to the plate and filament transformers. This cabinet is ventilated by means of intake louvers near the bottom, and an exhaust fan at the top. A pair of thermostats are visible near the top of each cabinet. One of each pair gives a high-temperature alarm at the terminal room, should excessive heating develop; the other removes the plate power, should a predetermined higher temperature be reached. In the upper half of the cabinet on the right the radio-frequency stages may be seen enclosed in a shielded compartment the cover of which has been removed. Plug-in crystal-oscillator chassis are at the left, the lower one being the spare unit. The box in the lower right corner of this compartment is the envelope demodulator which is placed directly below the modulated amplifier. A spare demodulator is provided with each transmitter and may be quickly inserted in place of the regular unit. The 4-stage signal amplifier is in the shielded compartment of several sections directly below the radio-frequency panel. On a half panel at the bottom of the cabinet on the right are mounted the relays and fuses, while on the left is the blower which provides a forced air draft on most of the tubes in this cabinet.

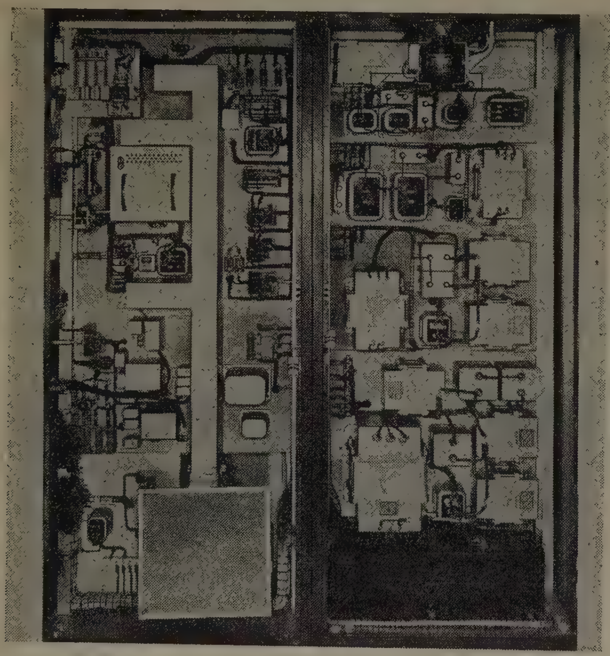


Fig. 3—Photograph of the right side of the transmitter.

The corresponding full right-side exposure of the transmitter is given in Fig. 3. A feature of interest in the left cabinet is the filtered-air intake for the blower and the vertical distribution duct through which the air is forced to the tubes. The air escapes through a thin spun-glass filter in the top of the cabinet. The transmitter monitor unit is the box similar to the demodulator which appears in the upper central position, and in the upper left corner the ground switch may be seen. The accessibility of apparatus in both cabinets is emphasized in this view.

The front view of the transmitter appears in Fig. 4. A particular advantage of this design is that all the tuning controls can be readily brought into a recessed compartment behind the front panel mat as shown. Meters are provided for measuring all significant voltages and currents through appropriate meter switches. Four lamps indicate the correct operation of the thermostats, the blower, and the holding-circuit controls. Four jacks on the front panel facilitate testing the signal-frequency-amplifier circuits. Six individual circuit breakers are provided, one for the primary power for each rectifier, one a master breaker for the four rectifiers, and the sixth is the main primary breaker for the transmitter. The latter removes all alternating-current power except that to the ovens of the crystal oscillators. The handle prominent in the upper right-hand corner operates the ground switch. In the ground position this switch opens the holding circuit, and directly grounds the high-voltage supply of each rectifier, and, through a mechanical linkage, releases the four identical door keys contained in the key barrels shown on the front panel. The switch is illuminated in the

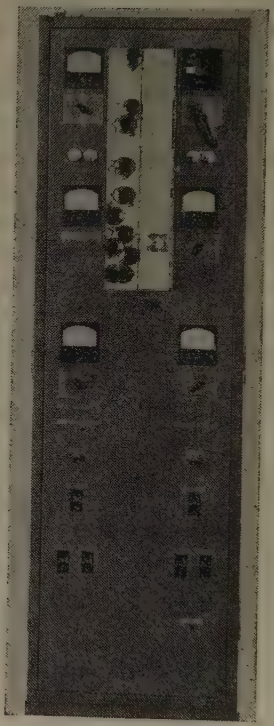


Fig. 4—Photograph of the front of the transmitter.

ground position and may be clearly seen through a meter bezel just above the switch handle. As a further safety precaution, opening any cabinet door also opens the high-voltage holding circuit. Remote control of the transmitter is effected through a relay, the contacts of which normally close the high-voltage holding circuit.

Circuit Description

Due to the stringent frequency-stability requirement imposed by the receiver design which is considered in detail in a companion paper,[2] the low-temperature coefficient quartz plate of the crystal oscillator is maintained at a nearly constant temperature. A fundamental crystal frequency of one twenty-fourth the final frequency is used, the second harmonic being selected in the oscillator plate circuit and amplified in the following stage. The power supply for these two stages is a 400-volt full-wave mercury-vapor rectifier.

The next two stages in the radio-frequency chain are single-tube harmonic generators which are operated as frequency doublers. The following stage is a two-tube harmonic generator operated as a frequency tripler to obtain the final frequency. A balancing control is provided to equalize the driving voltages applied to this unit. In the plate circuit a short lecher-frame system is used which is coupled to a similar lecher frame in the modulator grid circuit. Both lecher frames are tuned by external controls. The power for the three stages of harmonic generation, in which Western Electric 356A tubes are used, is obtained from a 1000-volt full-wave mercury-vapor rectifier.

The last stage in this chain is designated as the modulator. The signal voltage is introduced on the plates of the tubes of this push-pull amplifier which thus acts in accordance with the constant-current system of modulation. For this stage there was developed the Western Electric 364A tube which is similar in electrical properties to the 356A but provides two sets of independent plate and grid terminals.

A conventional parallel-tuned circuit is used in the first mesh of the modulator plate circuit. From each stator of the twin-stator plate condenser a short rod projects down to establish a small capacitative coupling with the demodulator pick-up probes, thereby supplying the radio-frequency envelope voltage to that unit. The coupling is so loose that the presence or absence of the demodulator unit is difficult to observe insofar as it affects the radio-frequency conditions of the modulator. The secondary mesh of the plate circuit is inductively coupled to the primary, and controls are provided for coupling in the load and varying the impedance into which the modulator operates. This two-mesh circuit is compactly mounted in turret style very close to the modulator tubes.

The first three stages of the type-K carrier-signal amplifier use low-power Western Electric D-159511 pentodes which are operated at a transconductance of approximately 10,000 micromhos. Stray capacitances are kept to a minimum in the design of the interstage circuits which are two-terminal networks the characteristics of which will be presented later in this paper. The power supply for these tubes and those in the monitor and demodulator is a 600-volt full-wave mercury-vapor rectifier. In each stage a series resistance drops the voltage developed on the plate to the right value. The correct screen potential is obtained from a potentiometer, while the bias required is obtained through a cathode resistance. The fourth stage is a Western Electric 363A, 350-watt power pentode, operating at a transconductance of approximately 7000 micromhos.

The modulation choke in stage four is a compact, low-capacitance, high-Q, air-core inductance wound in four sections and supported by an Isolantite tube. Screen and suppressor voltages as well as plate power for this stage are supplied from a 1250-volt well-filtered full-wave mercury-vapor rectifier. The frequency characteristic of the filter of this rectifier enters into the design of the gain-frequency characteristic of this stage at low frequencies. Direct-current plate power to the modulated amplifier is delivered through the modulation choke and a series voltage-dropping resistor suitably by-passed for the signal frequencies.

As shown in Fig. 1 a diode voltmeter is provided on the plate circuit of stage four to indicate the peak alternating-current signal voltage applied to the modulator. The gain of the four stages in tandem with 40 decibels of negative feedback is such that a single tone input at 28 decibels below 1 milliwatt will modulate the transmitter 80 per cent. The input impedance of the transmitter is 140 ohms. Without feedback the gain of these stages is approximately 110 decibels in the 12- to 60-kilocycle transmission band. To protect the signal amplifiers from excessive input signal voltage in case the feedback voltage is interrupted through a failure of the radio-frequency carrier, a relay is used in the grid circuit of the modulator which opens the primary circuit of the 600-volt rectifier when the grid current drops below a given value.

Test jacks appearing on the front of the transmitter enable the maintenance personnel to introduce a test signal from a standard signal generator on the grid of any stage. Provision is also made for the insertion of a shielded vacuum-tube voltmeter on the grid and on the plate of the stage under test. The grid of the following stage is disconnected and grounded, and the plate voltmeter capacity made equal to the grid capacity. This measuring technique allows the gain-frequency characteristic of each signal frequency amplifier, or of the four in cascade, to be readily obtained.

The demodulator is of particular interest, for it is from this unit that a facsimile of the radio-frequency envelope is obtained which is of correct phase to be reintroduced on the grid of the first stage of the signal amplifier

[2] D. M. Black, G. Rodwin, and W. T. Wintringham, "Ultra-short-wave receiver for the Cape Charles-Norfolk multiplex system," Proc. I.R.E., this issue, pp. 95–100.

to effect the negative-feedback correction of both distortion and noise. A departure from the conventional practice of using a simple diode rectifier or rectifiers is the use in this unit of symmetrical diode rectifiers each of which is acted upon independently by an amplifier driven by a small signal voltage developed across a resistance in the cathode of each diode. Each amplifier has approximately 25 decibels gain at signal frequencies. The plate of the amplifier is directly returned through radio-frequency filtering circuits and a direct-current blocking condenser to the anode of its driving diode. A marked improvement in the fidelity of the diode in rectifying the radio-frequency envelope at high per cent modulation is achieved by this local negative feedback connection on the diode. In addition, a remarkably flat transmission gain is realized out to approximately 2 megacycles although without the negative feedback the diode-cathode circuit has a cutoff frequency of approximately 75 kilocycles. A more comprehensive description of the functioning of this unit will be found in a companion paper.[3]

The transmission characteristic of the demodulator is of fundamental importance in considering the design requirements governing the amount of over-all negative feedback permissible without jeopardizing the stability of the transmitter. This characteristic is in general identified as a part of the β characteristic. Design features to achieve compactness, minimize stray capacitances, and to effect a high degree of isolation between the radio-frequency and the signal-frequency circuits are of great importance in obtaining trouble-free operation of this unit. It is appreciated that the inherent distortion in this unit limits the amount of the negative-feedback correction obtainable in the transmitter as a whole.

In the monitor two diodes are used, one connected directly to each side of the balanced coaxial transmission line which connects the transmitter to the antenna. A circuit similar to those of the demodulator employing one of the diodes as the radio-frequency envelope rectifier with a local negative-feedback amplifier is utilized as a means which, in conjunction with type K carrier terminal equipment, permits the measurement of the distortion characteristics of the transmitter. By inserting a vacuum-tube voltmeter across the plate circuit of the monitor amplifier, the signal-voltage amplitude due to modulation of the transmitter may be measured. This measurement, taken in relation to the radio-frequency carrier voltage indicated by the diode, allows the determination of the degree of modulation of the transmitter. It is readily shown that the voltage developed in the plate circuit of the monitor amplifier is within a few per cent of the actual envelope voltage when the gain of the amplifier is 25 decibels or more. The limitation on distortion measurements from the monitor is of the same order as the inherent distortion in the demodulator.

These transmitters may be operated into either an artificial antenna load or into the antenna array by means of a manually operated selector switch built into the junction box in which the gas-filled transmission lines supplying the antenna terminate. Two artificial antenna load resistors are placed in the junction box. From this box approximately 2 wavelengths of $\frac{7}{8}$-inch concentric-pipe transmission line are used to connect the selector switch to the output circuit of the modulator. The similarity of the load impedance presented by the antenna to that presented by the artificial-antenna-load resistors, and the degree of radio-frequency balance, may be noted by comparing the voltages indicated by the monitor diodes. The radio-frequency carrier power can be determined from the current of a thermocouple in series with the artificial antenna resistors of known value.

Circuit Characteristics and Performance Curves

In the paper by Burrows and Decino a comprehensive study of the design used in these transmitters is presented. In that paper the relationship between the gain-frequency characteristic of the transmitter and the stability of the transmitter with negative feedback is analyzed, and criteria established from which the component parts of the transmitter were then designed. The features of the negative-feedback design are clearly brought out by examining the $\mu\beta$ gain-frequency characteristic. This characteristic is the resultant of the gain-frequency characteristics of all the forward-acting, or μ circuits, plus all the gain-frequency characteristics of the backward-acting or β circuits.

Because of its importance, provision is made for measuring the $\mu\beta$ gain-frequency characteristic in the following manner. A known voltage from a variable-frequency standard-signal generator is applied to the grid of the third stage of the signal-frequency amplifier. To the plate circuit of the second stage of the signal-frequency amplifier, which for this test is disconnected from stage three, a shielded-vacuum-tube voltmeter is attached which simulates closely the circuit capacity normally present, and allows the measurement of the output voltage developed. In all other respects the transmitter remains normal. The ratio in decibel units of the output voltage of stage two to the applied voltage on stage three, as the frequency is varied, constitutes the $\mu\beta$ gain-frequency characteristic. This characteristic is plotted on semilog paper with frequency on the logarithmic scale to facilitate the recognition of the criteria established by W. H. Bode for negative feedback amplifier design.[4]

Since each interstage gain-frequency characteristic is a constituent of the $\mu\beta$ characteristic, the individual circuit characteristics must be carefully checked. Typical circuit characteristics are shown in Fig. 5. These characteristics were obtained by the use of auxiliary vacuum-

[3] C. R. Burrows, A. Decino, "Ultra-short-wave multiplex," Proc. I.R.E., this issue, pp. 84–94.

[4] W. H. Bode, "Relations between attenuation and phase in feedback amplifier design," *Bell Sys. Tech. Jour.*, vol. 19, pp. 421–454; July, 1940.

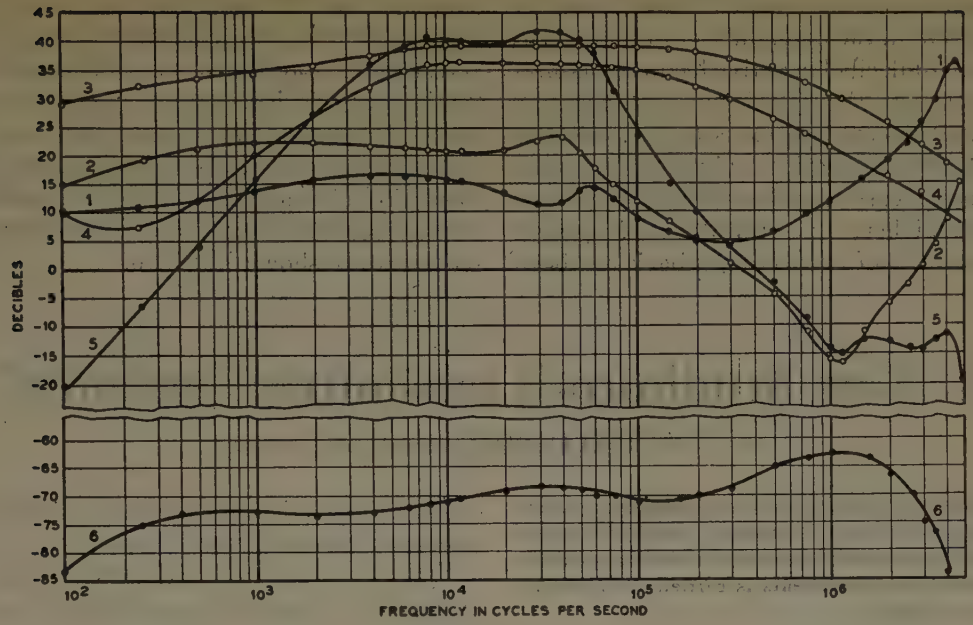


Fig. 5—Gain-frequency characteristic of the transmitter. The ordinates give the voltage gain in decibels plotted against the frequency on a logarithmic scale on the abscissa.

Curve (1) from the grid of Stage 1 to the grid of Stage 2
Curve (2) from the grid of Stage 2 to the grid of Stage 3
Curve (3) from the grid of Stage 3 to the grid of Stage 4
Curve (4) from the grid of Stage 4 to the plate of Stage 4 (Stage 4 load is the modulator)
Curve (5) from the grid of stage 3 to the plate of stage 2 (Normal $\mu\beta$ Characteristic)
Curve (6) The total β characteristic obtained from the sums of curves 1, 2, 3 and 4 less curve 5.

ube voltmeters and a variable-frequency standard-ignal generator. Such gain-frequency measurements xtend from approximately 25 cycles to as high as 30 negacycles, although the range of the $\mu\beta$ design may be overed adequately in going from 100 cycles to 5 mega-ycles. The difference between the sum of the gain-requency characteristics of the four forward-acting μ tages, and the over-all $\mu\beta$ characteristic, gives the gain-frequency characteristic. The β characteristic nay be further subdivided into the active elements haracteristic and the passive elements characteristic vhen the gain-frequency characteristics of the passive ircuit elements are known.

The radio-frequency operating characteristics of the nodulator which are of particular interest are the effec-iveness of the neutralization and the linearity between late voltage and load current. With respect to neutral-zation, an excellent condition was obtained with the imultaneous minimizing of the modulator plate cur-ents and maximizing of the grid currents. It was found hat a suitably linear relationship between the plate oltage and load current was realized for a load impe-lance permitting a carrier output power of approxi-nately 50 watts. For this condition the plate circuit andwidth, for a 3-decibel reduction in amplitude, is pproximately 1.8 megacycles.

By the simple expedient of grounding the negative-eedback return load, the transmitter may be operated vithout feedback although allowance must be made for a reduction in input voltage to the signal amplifier equal to the former amount of negative feedback. In this manner a set of distortion characteristics was obtained without benefit of negative-feedback correction. The feedback connection was then completed with negative feedback values of 30 and 40 decibels. No improvement in the distortion characteristics could be noted between the two values of feedback used, thus indicating that the

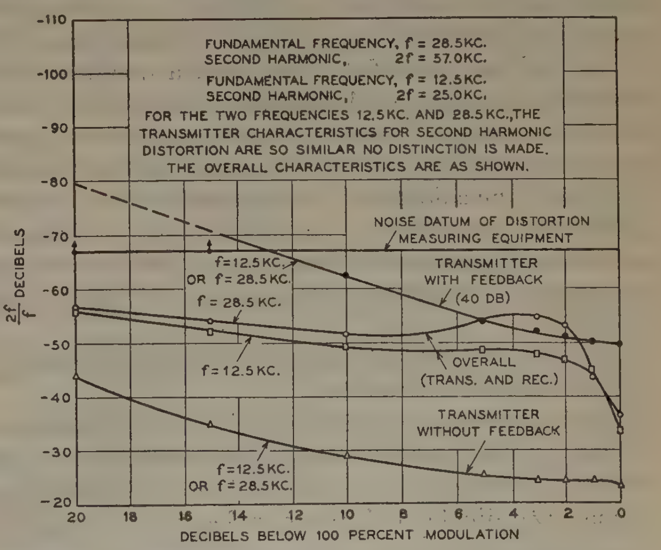


Fig. 6—Distortion characteristics of the transmitter and receiver with single-frequency input. The ordinates give in decibels the ratio of second-harmonic distortion to the fundamental as a function of the modulation level of the transmitter given in decibels below 100-per-cent modulation on the abscissa.

limitation in distortion measurement is to be found in the inherent distortion in the demodulator and monitor. However, the use of 40-decibel negative feedback assures the operating advantage that 10 decibels of negative feedback may be lost due to aging of tubes, etc., without penalizing the distortion correction realized in the transmitter. Furthermore, a corresponding reduction in noise is obtained with the higher value of feedback. Noise per channel with 40 decibels of negative feedback has been measured at 85 to 87 decibels below a signal which would modulate the transmitter 100 per cent. It is interesting to note that a negative-feedback voltage correction of 52 decibels could be used in these transmitters before any indication of instability was observed. This is in close agreement with design expectations. Typical distortion characteristics for the transmitter with and without feedback, and for the over-all radio system with feedback are shown in Fig. 6.

A New Studio-to-Transmitter Antenna*

M. W. SCHELDORF†, ASSOCIATE, I.R.E.

***Summary*—A directive antenna for studio-to-transmitter service is described, starting with a choice of structure based on physical requirements. A method for a general solution of radiation problems is described. Evolution of the components is carried out, and final performance curves are shown.**

Introduction

The antenna about to be described is one employed by the General Electric Company in its studio-to-transmitter link from the top of the International General Electric building at Schenectady to the main transmitter in the Helderberg Mountains, about 12 miles distant. The frequency is 343.6 megacycles. Fig. 1 is a view of the antenna from the street. It has been in continuous operation for two years, having passed through the severe sleet storm of the spring of 1943 without interruption of programs.

Choice of Structure

The Federal Communications Commission has set up requirements for the transmitting antenna for this service as follows, quoted from "Rules Governing S-T Broadcast Stations," section 4.34 (d) ". . . the gain in power toward the receiver shall be 10 (field gain 3.16) times the free space field gain from a doublet (137.6 millivolts per mile for 1 kilowatt at one mile). In all other directions 30 degrees or more off the line to receiver, the power gain shall not exceed $\frac{1}{4}$ the free-space field gain from a doublet." In graphical terms this means field-intensity limitations as shown in Fig. 2. Spherical co-ordinates are used for purposes of visualization and cartesian co-ordinates for purposes of engineering analysis.

The frequency range under the present allocations is 330.4 to 343.6 megacycles.

Radiating structures that will give the directivity required for this application fall approximately into three classes: (1) long wire systems such as the rhombic antenna; (2) dipole and plane reflector; (3) array of dipoles. The choice of one of these classes is principally a matter of physical requirements.

From a commercial viewpoint, an antenna for this application is preferably mounted on a single pole. This

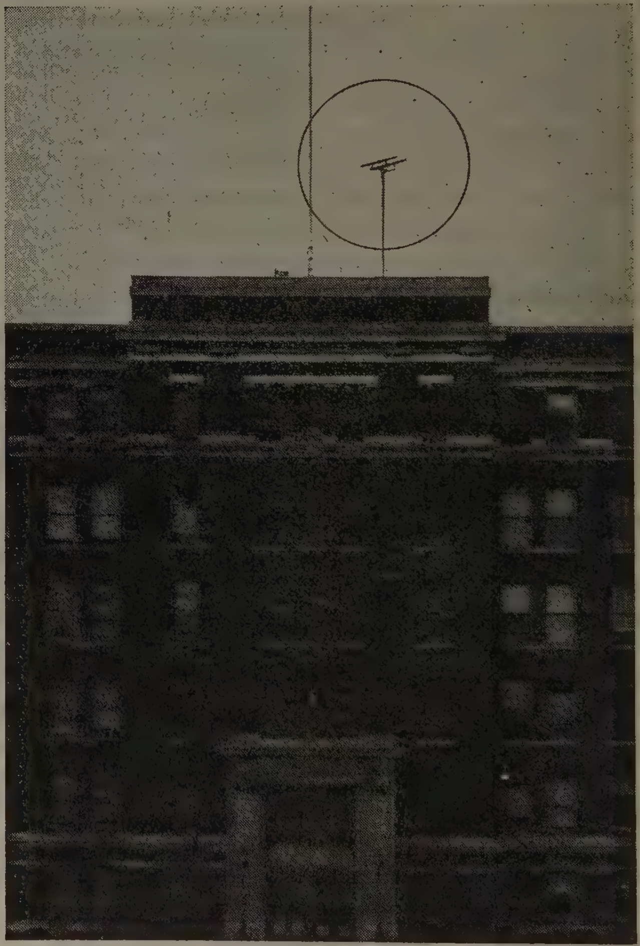


Fig. 1—General Electric radio antenna type MY36A atop building 36 of General Electric Schenectady works. View from street.

limits the dimensions in the horizontal plane and immediately eliminates class 1. The electrical system must be protected against ice and snow collection. This places a severe handicap on class 2. If the reflector is made of an open construction, ice and snow will fill in the spaces

* Decimal classification: R325.1×R550. Original manuscript received by the Institute, June 23, 1944; revised manuscript received, November 10, 1944.

† General Electric Company, Schenectady, New York.

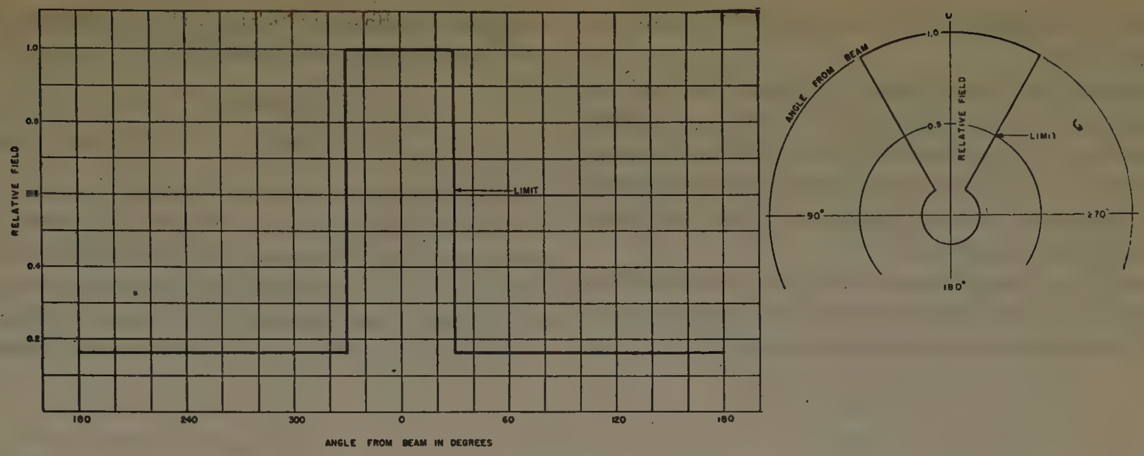


Fig. 2—Radiation limits.

and the wind load becomes as large as for a solid reflector. Due to the fact that a reflecting area must be provided both vertically and horizontally, this reflector load becomes very great, and the mounting arrangement becomes a problem. In the third class, it is possible to consider two linear arrangements of dipoles in a horizontal plane, each mounted in a cylindrical insulating tube which will protect the conductors from the weather and yet allow the radiation to take place with very little loss. In our case, we have chosen to place the entire radiating and connecting system within an airtight enclosure so that all metallic parts are pressurized and condensation problems are completely avoided.

Concentration of the radiation in the horizontal plane in the form of a single beam requires an array of radiators in the horizontal plane, and because of the simplicity of enclosing elements each of which lies in the horizontal plane as indicated above, the choice of polarization is automatically made. Obviously, the same beam can be secured from an array of vertical elements, but an enclosing structure would be difficult and the over-all structure would have a less acceptable appearance.

Theory

The determination of the disposition and number of radiating elements can be made in a rather systematic manner with the aid of some simplified relations concerning radiation from dipole arrays. It has been found[1] that the resultant radiation pattern for a complex arrangement can be secured by the simple product of terms representing the radiation from one dipole, with terms representing the disposition of the other dipoles, for extensions both in line and in parallel with the chosen basic dipole. For example, refer to Fig. 3. Here we have two sets of dipoles, each dipole l wavelength long, with m wavelength spacing in each set, and s wavelength spacing between the sets. The radiation in direction P is merely the product of the radiation from one of the dipoles A, for instance, multiplied by one factor representing the addition of one element B at a spacing m wavelength in line with it, and a second factor representing the addition of a second set of such dipoles A' and B' at a distance of s wavelength.

The casual reader, partly because of the length of the paper, may easily miss the fundamental importance of the relationships derived. In this paper we will endeavor to emphasize their usefulness and illustrate the application.

This multiplying procedure is limited to cases where all the elements are exactly the same and have the same magnitude and distribution of currents.

However, the phase of currents in group $A'B'$ can have any desired relation with respect to AB. The phase of currents in B can differ from A also, by any fixed amount so long as B' and A' differ by the same amount. In other words, a condition of symmetry must exist. If it is desired to have B different from A in current magnitude or distribution or both, but with B' different by the same manner from A', one may still apply a factor

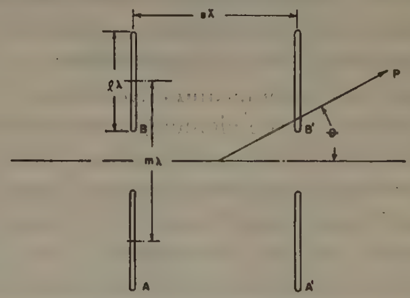


Fig. 3—General radiator array.

to the resultant radiation from set AB to get the over-all resultant. Stated simply, the procedure may be applied to the smallest common element to which the system may be broken down.

The resultant radiation in a vertical plane from a system such as that shown in Fig. 2 combined with several layers in the vertical direction (or several bays, as they are commonly called), can likewise be determined from the product of the radiation from one dipole, with one factor representing the addition of a set of dipoles spaced horizontally and with a second factor

[1] G. C. Southworth, "Certain factors affecting the gain of directive antennas," Proc. I.R.E., vol. 18, pp. 1502–1536, September, 1930.

representing the presence of a series of bays in the vertical direction.

It is obvious that this calculating procedure is of extreme value when one is interested in making adjustments in any of the parameters that are at his disposal. It is necessary simply to change the factor corresponding to the parameter change, the other factors and basic pattern remaining the same.

The use of the procedure outlined will now be illustrated, to show how we arrived at the fundamental radiating system of our antenna as shown in Fig. 4. As indicated, we have five half-wavelength dipoles end-to-end in each set and two sets spaced horizontally at three-fourths wavelength with a phase corresponding to this spacing. This spacing and phase may be recognized immediately by those familiar with radiation, as the necessary condition for complete cancellation of fields in the direction opposite to the main beam.

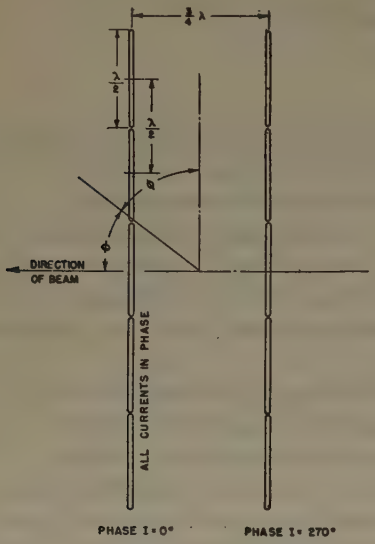


Fig. 4—Chosen radiator composition.

To simplify calculations and adjustment of the array, we chose to keep all the elements alike with respect to size, current magnitude, and current distribution, so that the smallest common element was the half-wave dipole. Its radiation pattern is shown in Fig. 5, which shape is exactly described by the expression $E=K\cos(\pi/2\sin\theta)/\cos\theta$, a slight variation from the common shape often shown, i.e., two tangent circles.

The general expression for factors due to combinations of antennas is taken from Southworth[1]

$$f(\psi) = \frac{\sin n\pi(a\cos\psi + b)}{n\sin\pi(a\cos\psi + b)}.$$

Where n is the number of common elements, a is a numeric equal to the center-to-center spacing between adjacent elements in wavelengths, ψ is the angular deviation from a line through the centers of the elements in the plane desired, and b is the time-phase angle between currents in adjacent elements.

Applying this to two dipoles placed end to end, with currents in phase, we obtain the following: $a=1/2$ and $b=0$ and the expression for the net radiation is

$$F = K\left[\frac{\cos(\pi/2\sin\theta)}{\cos\theta}\right]\left[\frac{\sin(\pi\cos\phi)}{2\sin(\pi/2\cos\phi)}\right].$$

With reference to the two angles given in this expression, see Figs. 4 and 5. The first part uses an angle θ measured from the normal to the elements, as shown in both figures. The second part must use an angle measured from a line through the centers of the elements, which would be $90°-\theta$ or ϕ as we have shown.

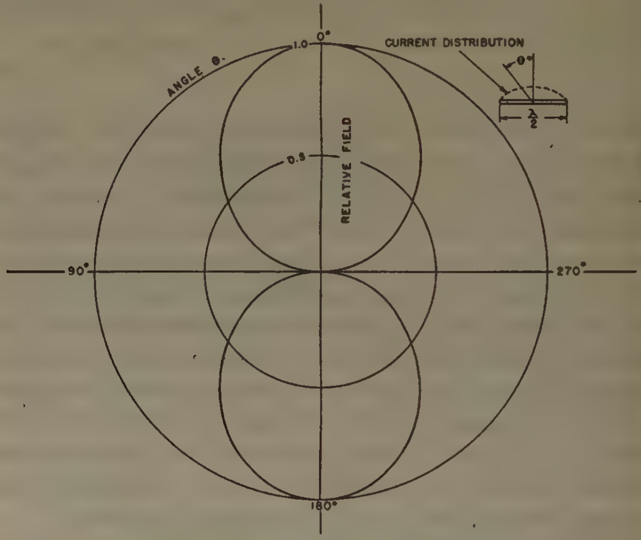


Fig. 5—Radiation from half-wave dipole.

For additional dipole elements, the first factor of the above expression remains the same, and the second is altered slightly. Fig. 6 shows the factors $f(\phi)$ forming the second part of the expression for dipole combinations up to six elements end-to-end, and Fig. 7 shows the net radiation of these antenna combinations. The heavy lines in Fig. 7 show the superimposed Federal Communications Commission limits, indicating that four elements are essentially sufficient, if it is possible by other means to reduce the back radiation to a sufficiently low value.

It is necessary next to study the effect of horizontal sets of elements, in reduction of this back radiation.

In the first place, we have found by experience that parasitic reflectors cannot be used to control the radiation as satisfactorily as is possible with directly-excited reflectors. This implies that both sets of elements will be energized by some form of transmission-line system. The most direct method is to use a feed line between the sets, of a length equal to the spacing, which for practical purposes fixes the time-phase-angle difference at a value equal to the space-phase angle. For simplicity, n is kept at the value 2. Fig. 8 shows the factors $f(\theta)$ representing the effect on a radiation pattern by sets of elements in a plane spaced at distances from one-eighth wavelength to a full wavelength.

It is obvious that we cannot use any spacing other

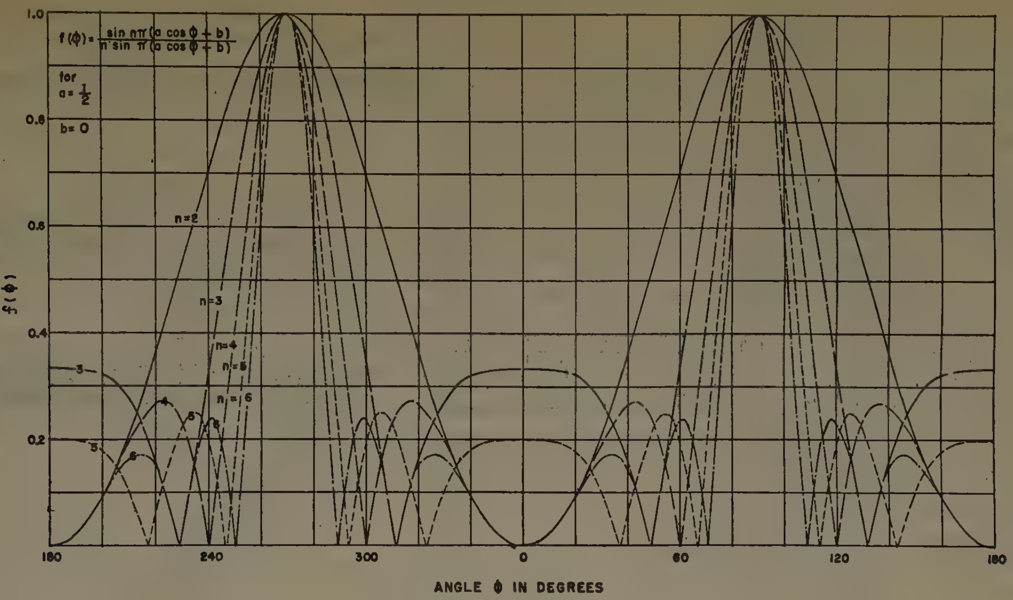


Fig. 6—Radiation factor.

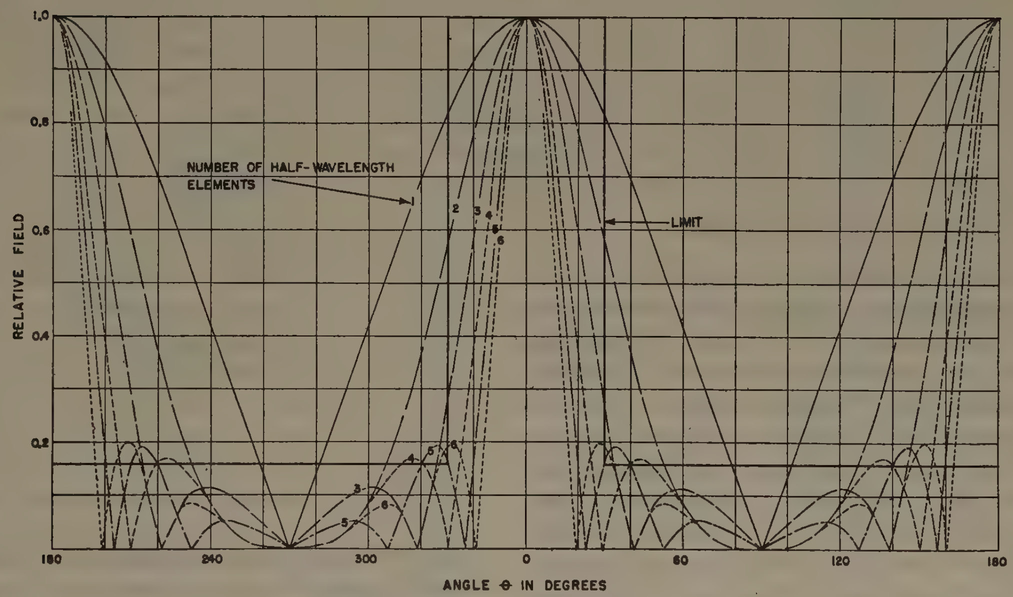


Fig. 7—Radiation from dipoles in line.

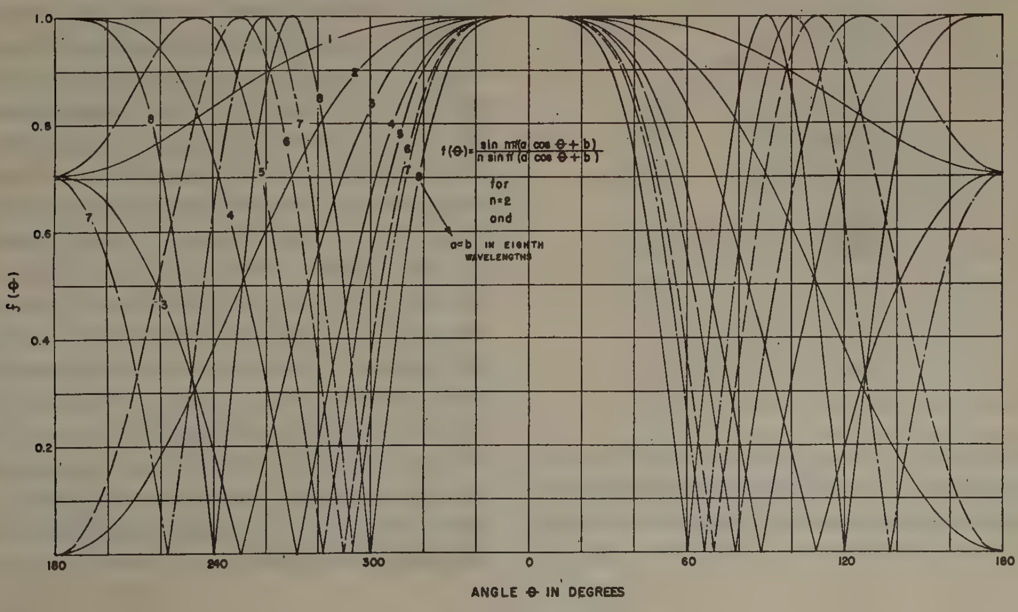


Fig. 8—Radiation factor.

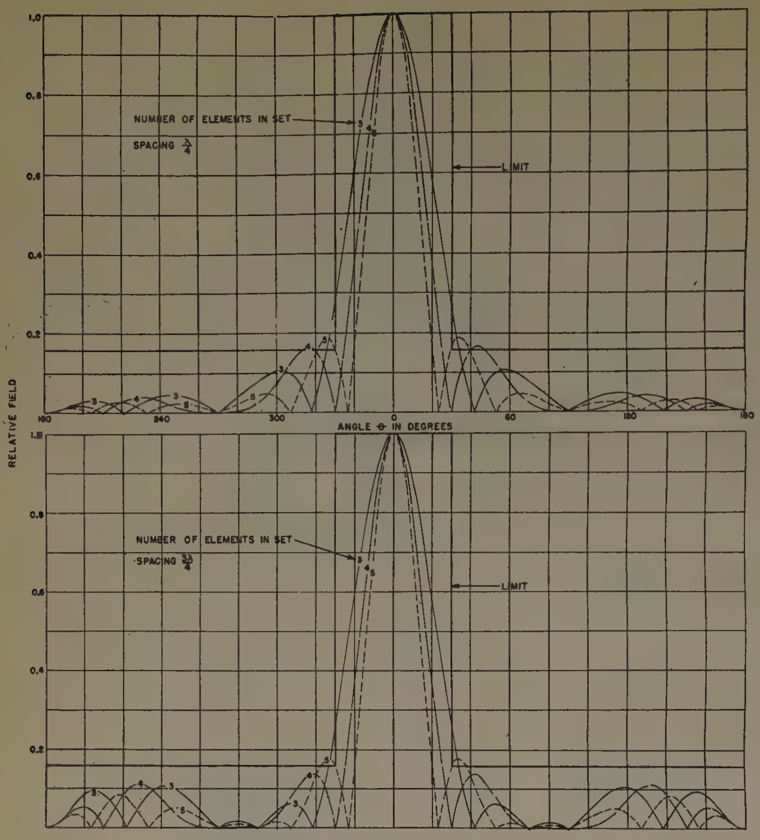


Fig. 9—Radiation from sets of dipoles.

than substantially one-fourth or three-fourths wavelength, and secure sufficient reduction of the back wave.

Fig. 9 shows the application of the corresponding factors to the sets of elements in line, for element quantities of 3, 4 and 5 in each set. The greater spacing shows a definite improvement in the first side lobe which is forward, with a permissible increase of side lobes in the rear. In addition, this spacing has the advantage that approximately one third the mutual impedance will be experienced between sets, due to the greater spacing. This is especially important in the case of antennas

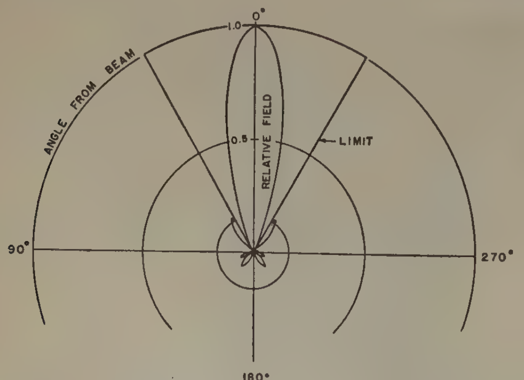


Fig. 10—Theoretical radiation for five-element array.

having linear dimensions of several wavelengths such that mutual impedances may become accumulative. Having chosen the spacing, only one factor remains to be settled, that of the number of elements in each set.

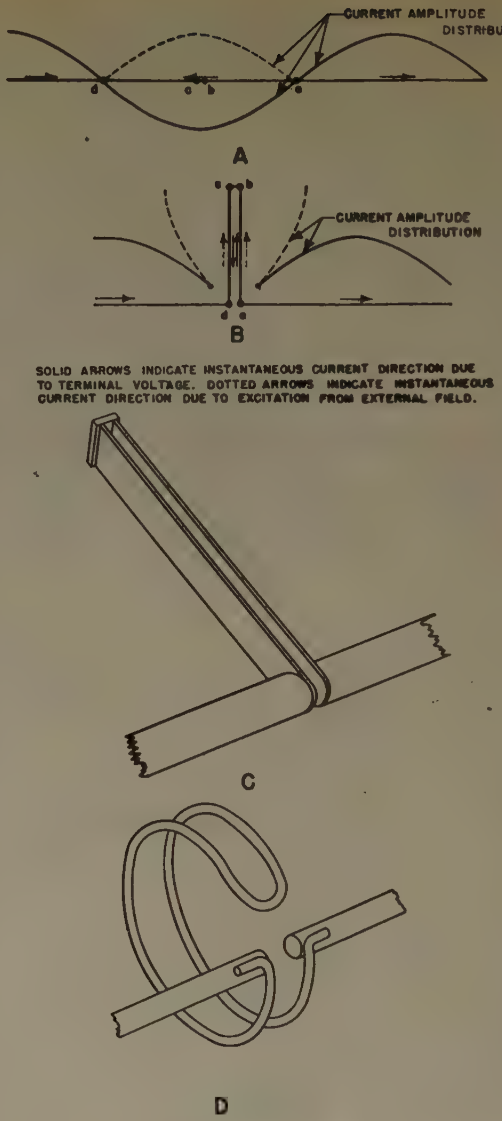


Fig. 11—Phase inverters.

Experimentally, we have found that the main beam is usually wider than that called for by theory. It is consequently improbable that three elements in a set could ever be made to meet the requirements. Four elements in each set are theoretically sufficient, but this condition has a decided disadvantage. It requires (if we wish to have but one feed point in each set of conductors) that the system be end-fed, that is, at high terminal impedance, which is well known to give shunt-capacity trouble and results in an unnecessarily high matching ratio. Also in the physical structure which we planned for this antenna, we had a large tubular metal support placed horizontally at the center, between the sets of elements, and considerable metallic mass at the ends of this support. This, we anticipated, would upset the radiation pattern, and we felt it would be desirable to have the theoretical beam well within the limits. For this reason the five-element array was chosen. It is true that the first side lobe theoretically fails to pass, but we have found experimentally that this is not the limitation, as will be borne out by the final performance curves

Plotted on Fig. 10 is the chosen theoretical curve, in ɔlar form.

The theoretical power gain for this form is 14.5, well excess of the 10 required.

Practice

The problem of putting this antenna into a practical ɾm for manufacture, involved principally the realiza- on of equal currents in all the elements, with a time- hase-angle difference of 270 degrees between the sets of ements. As indicated previously, for practical reasons is desirable to feed each set of elements at only one ɔint, which, because of symmetry requirements, places ɩis point at the center of the middle element, at a low npedance level, which is a preferred condition. Excita- on of the other elements is made by means of phase- ıverting elements as illustrated in Fig. 11B.

This is a very common method and has been very ex- ɛnsively applied by amateurs.[2] The behavior may be nderstood by comparison of A and B in Fig. 11. shows the condition of standing waves of current set ɔ in a wire in free space, when the left-hand end is ɛcited at a constant frequency. Note that at points and d, where the current amplitudes go through zero, ıe instantaneous directions of current are opposite. his makes it possible to fold the conductor which has ıe reverse current at the points a and d to secure the ɔnfiguration of B. Here the instantaneous currents are ɔposite in direction and equal in magnitude at all posi- ons so that there is no field produced by the current ɔw. The net result is currents in successive half-wave- ngth sections which are flowing in the same direction is required for successful array operation.

We incorporated this plan into our first antenna as dicated by the construction of C in the figure, but ter considerable experimental procedure were able to ıow that this particular interpretation of the theory ıs decided limitations. Due to a definite measured field ɔm the antenna system in line with the conductors, ɛ., $\theta = 90$ and 270 degrees (where there should be none), e established the fact that the phase-inverting ele- ents were acting as radiators. This is a result of voltage duced in them by currents in the antenna elements. hese voltages are such as to produce current in the ɔo branches in the same direction as shown by the ɔtted arrows in B, Fig. 11. A solution for the trouble shown in D, where it is evident that cancellation of ternal fields from the inverter itself is essentially fected.

The disturbance of the pattern by currents in the nple inverters is also found in conductors and metallic pports which lie in the plane of the sets of elements. ıerefore, it was necessary to support each set of ele- ents in the vertical direction and to bring the feeding ansmission lines to the terminals in the same general rection. This accounts for the rather odd physical

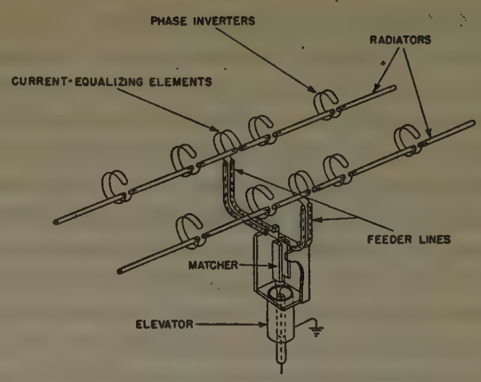


Fig. 12—Electrical system of antenna.

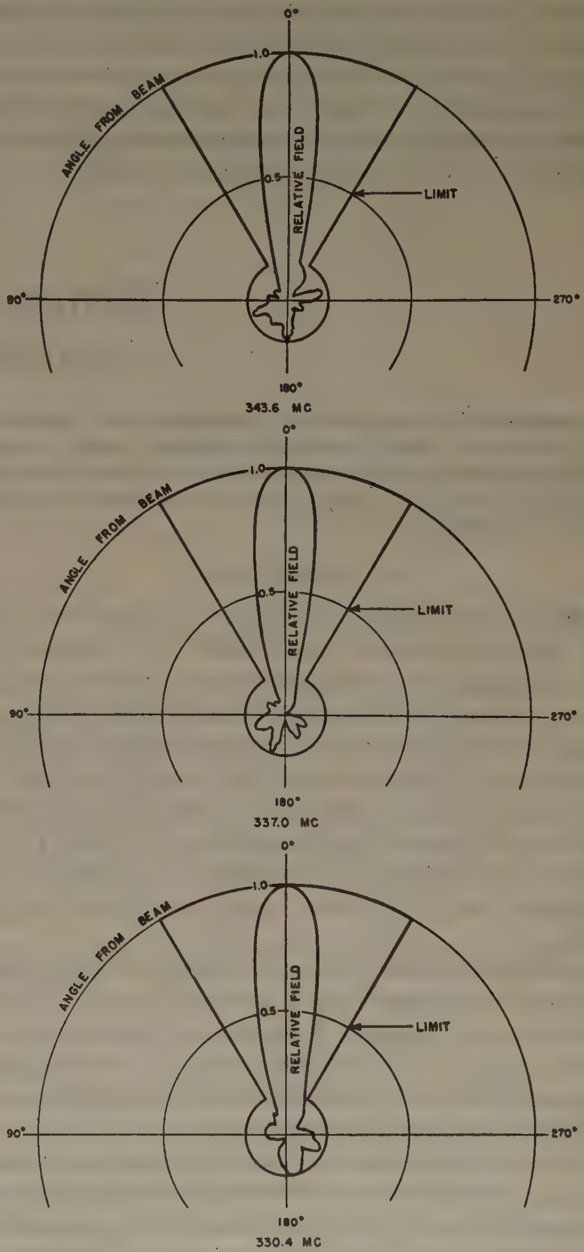


Fig. 13—Experimental radiation curves.

structure at which we arrived. The diagram of Fig. 12 shows the entire electrical system. Starting from the main transmission line, we have first an elevator or line-balance converter, to permit attachment to a balanced

[2] "Antenna Book," published by the American Radio Relay ague, West Hartford, Connecticut, no. 15, pp. 58–60; September, 39.

load. Next to it is an adjustable matcher. The feed lines are different in length by one-quarter wavelength, a convenient length made use of by the fact that the junction point, which is at the center of a junction box, is of necessity on one side of the antenna mounting pole. This difference in lengths achieves the same purpose as a three-fourths-wavelength difference, except that the beam from the antenna is reversed. Between each feeder line and the antenna terminals is connected an adjustable impedance line which acts as an impedance converter, not really a matcher because its electrical length is not a quarter wavelength. It is not a desirable element, but we found that we were unable to get the necessary pattern without it, that is, it was necessary in order to achieve equality of current in the two sets of elements. A number of alternative solutions were tested but consistently we found that the extensive mutual impedance between the sets of elements caused the currents in them to increase or decrease together. It was relatively easy to secure very low back radiation with large side lobes or very low side lobes with a large back lobe, but never the desired combination, unless the current-equalizing elements were used.

The radiation patterns over the frequency band are shown in Fig. 13. There was no change in the physical structure for these curves, which is highly desirable from a commercial standpoint. We were unable to verify quantitatively the calculated gain, but the shape of the main beam of the radiation pattern and the magnitude of the side lobes assure safe compliance with the mini mum value required.

ACKNOWLEDGMENT

We are indebted to Mr. R. C. Longfellow, of the General Electric Company, for his work on theoretical methods of antenna analysis similar to those used herein and to Messrs. J. E. Brown and R. W. Hodgers for their assistance in the preparation of this paper.

Reflex Oscillators*

J. R. PIERCE†, ASSOCIATE, I.R.E.

Summary—**This paper discusses qualitatively the behavior of reflex oscillators. Power production, electronic tuning, variation of frequency with resonator voltage, effect of modulation coefficient, and influence of load are considered. Two brief mathematical appendixes are included.**

I. INTRODUCTION

THE reflex oscillator is a form of high-frequency long-transit-time tube which has distinct advantages as a low-power source. It may be light in weight, need have no magnetic focusing field, and may be made to operate at comparatively low voltages. The reflex oscillator may serve as a beating oscillator in double-detection receivers or as a frequency-modulated oscillator in low-power transmitters. So far the efficiencies which have been attained are quite low, but that need not be a severe handicap in some applications.

In this paper some general aspects of reflex oscillators will be discussed in a qualitative manner with the aid of various diagrams. It is believed that such a discussion may be of considerable value in orienting one's thinking about a type of tube which may be comparatively new to many radio engineers.

Although this is intended to be a descriptive rather than a mathematical paper, two brief appendixes are attached indicating the derivation of some expressions used, and mathematical expressions for the various curves used have been indicated in the figures. The symbols in the mathematical expression of the figures are defined in the appendixes. Readers are referred also to early papers on allied work by several writers.[1–7]

II. GENERAL DESCRIPTION

Reflex oscillators can be considered as oscillators in which an electron stream passes through a longitudinal radio-frequency field across a "gap" between two electrodes, then into a drift space in which there is a retarding electric field produced by a negative repeller electrode, and finally returns through the radio-frequency field across the gap.

Reflex oscillators are not entirely new. Fig. 1 shows an early "reflex" oscillator, the negative-plate Barkhausen tube operated with the resonant circuit between cathode and grid. In this tube, the initial radio-frequency field through which the electron stream passes is applied across the "gap" between the cathode and the positive grid. The drift space with its retarding field is between the grid and the negative plate. This retarding field returns the electrons through the gr

* Decimal classification: R339×R113. Original manuscript received by the Institute, May 26, 1944; revised manuscript received, August 21, 1944. Presented, Winter Technical Meeting, New York, N. Y., January 26, 1945.
† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] A. Arsenjewa-Heil and O. Heil, "Eine neue Methode zur Erzeugung kurzer, ungedämpfter, elektromagnetischer Wellen grosser Intensität," *Zeit. für Phys.*, vol. 95, pp. 752–762; July, 19
[2] Russel H. Varian and Sigurd F. Varian, "A high frequency oscillator and amplifier," *Jour. Appl. Phys.*, vol. 10, pp. 321–327; May, 1939.
[3] W. C. Hahn and G. F. Metcalf, "Velocity-modulated tubes," PROC. I.R.E., vol. 27, pp. 106–116; February, 1939.
[4] David L. Webster, "Cathode ray bunching," *Jour. Appl. Phys.*, vol. 10, pp. 501–508; July, 1939.
[5] W. C. Hahn, "Small signal theory of velocity-modulated electron beams," *Gen. Elec. Rev.*, vol. 42, pp. 258–270; June, 1939.
[6] Simon Ramo, "The electronic-wave theory of velocity-modulation tubes," PROC. I.R.E., vol. 27, pp. 757–763; December, 1939.
[7] "Klystron Technical Manual," Sperry Gyroscope Company, Inc., Brooklyn, N. Y., 1944.

across the "gap" between the grid and the cathode. This form of reflex oscillator has the disadvantage that electrons which lose energy in the first crossing of the gap cannot again cross the gap and reach the cathode under a retarding radio-frequency field. The disadvantage is not fatal. At high frequencies reflected elec-

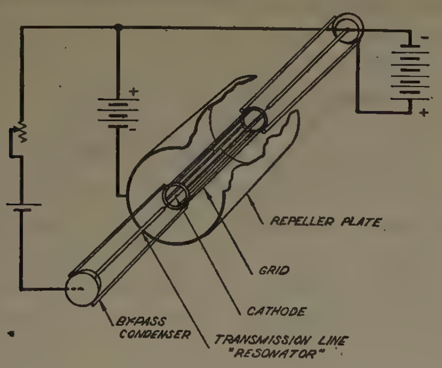


Fig. 1—An early reflex oscillator:—the negative-plate Barkhausen tube.

trons, which merely penetrate into the cathode-grid region and are turned away from the cathode without reaching it, can give up energy to the circuit connecting the cathode and grid.

Although reflex oscillators themselves are not entirely new, there is a new terminology, that of velocity modulation, drift action, and bunching, which makes it possible to explain the action of reflex oscillators in simple terms. Fig. 2 explains the principle of the drift space and bunching in a retarding field. The retarding field may be, for example, the gravitational field of the earth. The drift time may be the time required by a ball thrown upward to return. If the ball is thrown upward

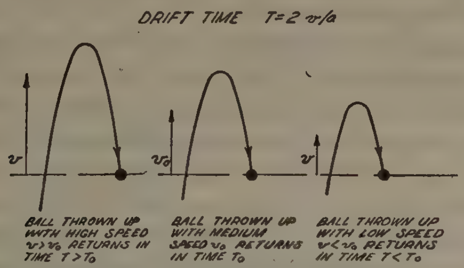


Fig. 2—Principle of the drift space.

with some medium speed v_0, it will return in some time T_0. If it is thrown upward with a low speed v smaller than v_0, the ball will return in some time T smaller than T_0. If the ball is thrown up with a speed v greater than v_0, it returns in some time T greater than T_0. Now imagine three balls thrown upward in succession, evenly spaced in time but with large, medium, and small velocities, respectively. As the ball first thrown up takes a longer time to return than the second, and the third takes a shorter time to return than the second, when the balls return they will be closer together than when they were thrown upward. Thus "bunching" occurs when the velocity with which a uniform stream of particles is projected into a retarding field is progressively decreased.

Fig. 3 shows a modern reflex oscillator. In this oscillator the gap is defined by two grids, at a positive potential with respect to the cathode, which accelerate the electrons to a sufficient speed so that even electrons which have lost a certain amount of energy to the radio-frequency field on the first crossing of the gap can still re-cross the gap against a retarding radio-frequency field. By referring either to Fig. 3 or Fig. 1, one can see how velocity modulation and drift action are utilized in producing radio-frequency power. The electrons leave the cathode and enter the radio-frequency field across the gap in a uniform stream and there receive a velocity modulation. In drifting in the retarding field produced by the repeller and returning to the gap, the electrons which passed across the gap when the field was becoming progressively less accelerating became bunched, and the electrons which passed across the gap when the field was becoming progressively more accelerating be-

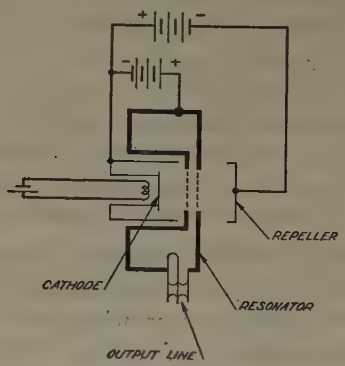


Fig. 3—A modern reflex oscillator.

came spread out. Thus the returning electron stream forms a pulsating current when it again crosses the gap. By thinking through the drift action, remembering that the current induced by a given electron in the circuit flows in opposite directions for the two successive transits across the radio-frequency field in the gap, one can see that the electron stream will give power to the field across the gap when the time between leaving the field and returning across it is $n+\frac{3}{4}$ cycles, where n is any integer, including zero.

III. Power Production

Important in understanding the functioning of the reflex oscillator is the dependence of power upon length of drift time, voltage across the gap, and other parameters. It is rather obvious that the power output will increase as the direct-current voltages and currents are increased. The other important factors in determining power output are drift time, (which will be measured in number of cycles, N) and radio-frequency voltage across the gap, which is controlled by the loading of the oscillator. Fig. 4 shows plots of radio-frequency electron current in the electron stream returning across the gap

and radio-frequency voltage across the gap.[8] As might be expected, the greater the number of cycles the electrons drift in the drift space, the lower the radio-frequency gap voltage required to produce a given amount of bunching and hence a given radio-frequency electron current. It may be seen from Fig. 4 that as the radio-frequency gap voltage is increased, the radio-frequency electron current gradually increases until a maximum value is reached, representing as complete bunching as is possible, after which the current decreases with increasing gap voltage. The maximum value of the current is approximately the same for various drift times, but occurs at smaller gap voltages for longer drift times.

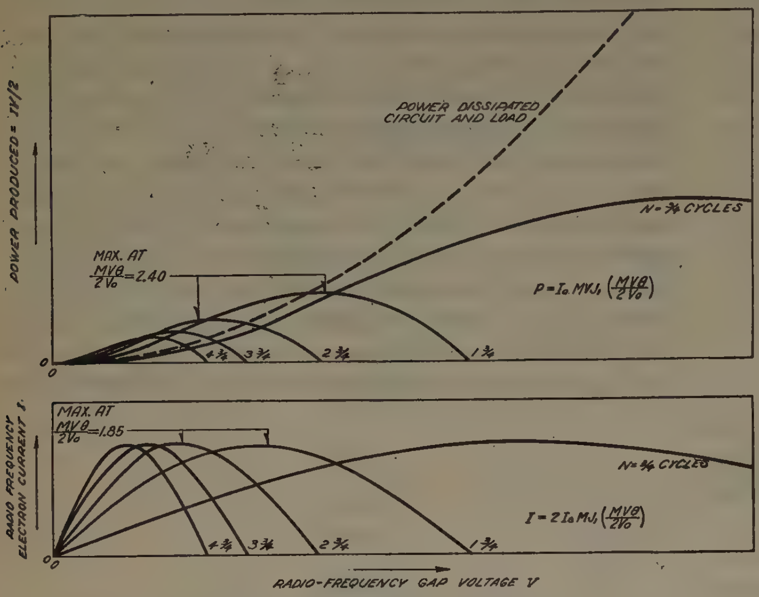


Fig. 4—How current and power vary with gap voltage and with number of cycles drift N.

The radio-frequency power produced is the voltage times the current. As the given maximum current occurs at higher voltages for shorter drift times, the maximum power produced will be greater for shorter drift times. This is clearly brought out in the plots for power versus voltage shown in Fig. 4.

The power dissipated in the circuits and load will vary as the square of the radio-frequency voltage. Part of this power will go into the load coupled with the circuit, and part will be put into unavoidable circuit losses. A typical curve of power into the circuit and load versus radio-frequency voltage is shown in Fig. 4. Steady oscillation will take place at the voltage for which the power-production curve crosses the power-dissipation curve. For instance, in Fig. 4 the power-dissipation curve crosses the power-production curve for $1\frac{3}{4}$ cycles drift at the maximum or hump of the curve. This means that the circuit impedance for the dissipation curve shown is such as to result in maximum production of power for $1\frac{3}{4}$ cycles drift. For $2\frac{3}{4}$ cycles drift and for longer drifts, the power-dissipation curve crosses the power-production curves to the right of the maximum, and hence the particular circuit loading shown does not result in maximum power production for these longer drift times. This is an example of operation with lighter-than-optimum load. The power-dissipation curve might cross the power-production curve to the left of the maximum, representing a condition of too heavy loading for production of maximum power output. The power-dissipation curve in Fig. 4 always lies above the power-production curve for a drift of $\frac{3}{4}$ cycle. This means that the tube for which the curves are drawn loaded to give the power-dissipation curve shown could not oscillate with the short drift time of $\frac{3}{4}$ cycle, corresponding to a very negative repeller voltage.

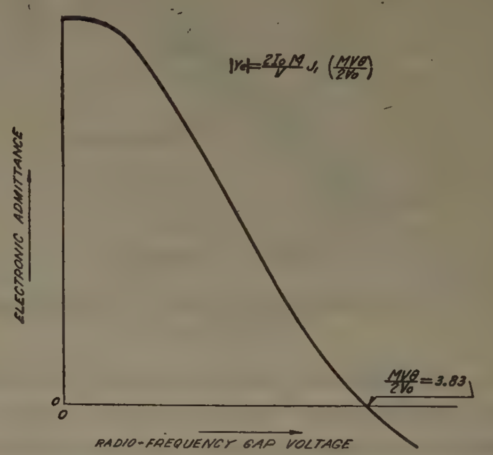


Fig. 5—How electronic admittance decreases as amplitude of oscillation increases.

In general, the conclusions reached by examining Fig. 4 are borne out in practice. The longer the drift time, that is, the less negative the repeller, the lower the power output. For very-negative repeller voltages, however, corresponding to very short drift times, the power either falls off, which means that most available power is dissipated in circuit losses, or the tube fails to operate at all because, for all gap voltages, the power dissipated in circuit losses is greater than power produced by the electron stream.

IV. Electronic Tuning

An important aspect of reflex oscillators is the possibility of tuning them electronically. The frequency of oscillation can be changed by a substantial amount, usually several tens of megacycles, by varying the voltage of the repeller electrode. Both the amount of electronic tuning which can be obtained and the rate of change of frequency with voltage can be explained simply in terms of admittance diagrams.

Because of velocity modulation and drift action, the electron stream produces an admittance across the gap. Because of nonlinearities, for a given drift time this electronic admittance decreases in magnitude as the

[8] See Appendix A for origin of curves. These curves are based on an approximation and are not valid for large values of V. Thus the curves for $N=\frac{3}{4}$ cycle are in error in the "maximum-power" region.

radio-frequency gap voltage is increased. This can be seen by examining the curves for current versus voltage shown in Fig. 4. The electronic admittance for any voltage may be obtained in dividing the radio-frequency electron current by the radio-frequency gap voltage. Fig. 5 shows how the magnitude of electronic admittance varies with radio-frequency gap voltage. Fortunately, explanations of the behavior of reflex oscillators are greatly simplified by the fact that the phase of the electronic admittance is unaffected by the radio-frequency gap voltage.

Fig. 6 shows a plot of electronic admittance produced across the gap by the electron stream for very small radio-frequency voltages, plotted as a function of drift angle θ, which is 2π times the number of cycles drift, N. This small-signal electronic-admittance plot takes the form of a spiral, in which the amplitude gradually increases, and the phase continually changes as θ is increased. The crossing of the negative-conductance axis first from the origin corresponds to $\frac{3}{4}$ cycle drift, the second crossing to $1\frac{3}{4}$ cycles drift, etc. For a larger radio-frequency gap voltage the variation of electronic admittance with θ may be represented by a similar spiral with all radii reduced as indicated in Fig. 5. Also shown in Fig. 6 is minus-the-admittance of a tuned circuit near resonance. This admittance plot takes the form of a straight vertical line. The distance from the origin along a horizontal axis is minus-the-conductance of the circuit, which near resonance does not change appreciably with frequency. Susceptance, measured in the vertical direction, is proportional to frequency off resonance and to the effective capacitance of the resonant circuit.

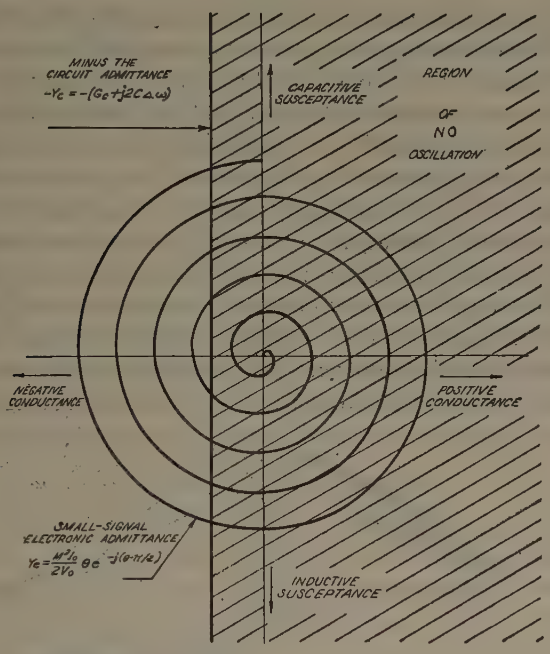


Fig. 6—Condition for oscillation.

In order for oscillations to build up the electronic conductance, that is, the horizontal component of the electronic admittance, oscillations must be negative and have a magnitude greater than the circuit conductance. This means that oscillations can occur only in the region of the admittance diagram not cross-hatched in Fig. 6.

The electronic-tuning range between extreme frequencies which can be obtained by changing θ in the neighborhood of one optimum-power value is proportional to the vertical distance between intersections of the spiral and the minus circuit-admittance line. It will be observed in examining Fig. 6 that when the circuit conductance is small enough so that the electronic-admittance spiral cuts the circuit-admittance curve near the vertical axis, changes in circuit conductance will cause only small changes in the vertical position of intersection and hence will cause only small changes in the electronic-tuning range. When the circuit conductance is almost as large as the electronic admittance for small gap voltages, changes in circuit conductance may, however, produce large changes in the electronic-tuning range.

For steady oscillation, the electronic admittance must, of course, be minus the circuit admittance. As oscillations build up, the electronic-admittance spiral shrinks in accordance with Fig. 5 until this condition is fulfilled. Thus, from curves such as those shown in Figs. 5 and 6, the gap voltage or amplitude of oscillations and the frequency of oscillation can be obtained for any drift angle θ.

It will be observed that the electronic-admittance spiral crosses the circuit-admittance line at points farther from the horizontal axis for longer drift times. This means that the longer the drift time, the farther off circuit resonance the oscillator can be made to oscillate, and, hence the greater the range of electronic tuning which is available. The diameter of the admittance spiral increases with the increasing direct-current and with decreasing direct voltage, as well as with increasing drift time. As has already been pointed out, the circuit susceptance varies more rapidly with frequency as effective circuit capacitance is increased. Thus we may conclude that the greater the direct beam current, the lower the direct voltage at the gap (disregarding effects on the modulation coefficient, which will be discussed later), the longer the drift time, and the smaller the effective capacitance of the circuit, the larger will be the range of electronic tuning. This is borne out in practice, and it can be particularly noted that the electronic-tuning range is greater for lower repeller voltages, that is, for longer drift times.

The diagram in Fig. 7 explains why frequency changes more rapidly with drift angle as the oscillator is loaded more heavily, thus reducing circuit Q. In this diagram, part of the electronic-admittance spiral and two circuit-admittance curves are shown. One of these latter is close to the vertical axis, corresponding to light loading of the oscillator; the other is farther from the axis, corresponding to heavy loading of the oscillator. Height above the horizontal (zero susceptance) axis along either of the minus circuit-admittance curves is proportional

to frequency off circuit resonance. The angle of a line from the origin (zero susceptance point) through a point on a minus circuit-admittance curve gives the phase of minus-the-circuit-admittance for that point, and hence gives the phase the electronic admittance must have for oscillations to occur at the frequency corresponding to

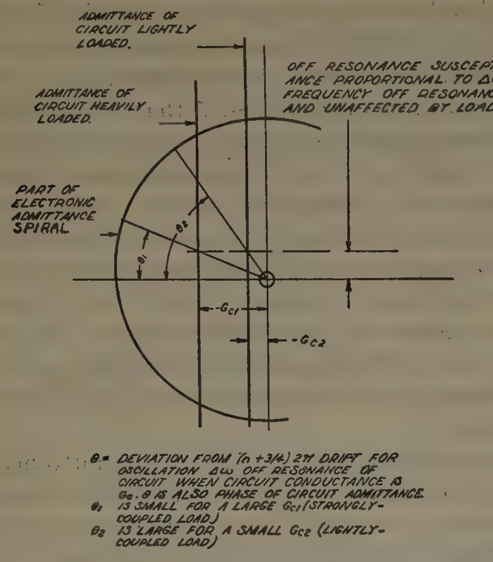


Fig. 7—Why frequency changes more rapidly with drift angle as the oscillator is loaded more heavily (thus reducing circuit Q).

that point. As oscillations build up the electronic-admittance spiral shrinks, as has been explained, but its phase will not change. It may be seen that, for a given frequency off resonance, the required electronic-admittance angle with respect to that for $n+\frac{3}{4}$ cycle drift will be small for the large-circuit conductance, corresponding to heavy loading, and large for the small-circuit conductance corresponding to light loading. This angle is controlled by varying the repeller voltage. In order for the oscillator to operate off resonance by a given frequency, the phase of the electronic admittance, and hence the drift angle, must be changed more in the case of a lightly loaded circuit than in the case of a heavily loaded circuit, and in order to change the phase more, the repeller voltage must be changed more. This is a characteristic of reflex oscillators that is borne out in practice.

From figures such as 5, 6 and 7, and from the relation between drift time and repeller voltage, curves of amplitude versus frequency and of frequency versus repeller voltage can be derived. Fig. 8 shows such curves for several load conditions. Actual behavior of reflex oscillators is very similar to that illustrated by these. Of course, circuit losses make the cusp-like amplitude-versus-frequency curve for zero load unattainable, but a sharpening of the top of amplitude-versus-frequency curves may be noted at low loads. It should be noted that in these amplitude-versus-frequency curves, relative amplitude is shown. Of course the actual radio-frequency voltage will be high for light loading and low for heavy loading. The S shape of the frequency-versus-repeller-voltage curves is very noticeable in practice.

V. Variation of Frequency with Resonator Voltage

The fundamental variable in electronic tuning is drift time. Drift time is changed by changing voltage. So far, the variation of frequency with repeller voltage has been discussed. The frequency of operation is also a function of the direct voltage between the cathode and the gap in the resonator across which the electrons travel, here called the resonator voltage. Making the repeller more negative invariably shortens the drift time and increases the frequency. Increasing the resonator voltage may, however, either shorten or lengthen the drift time, and change the frequency in either sense. Increasing the resonator voltage increases the mean speed of the electrons in the drift space, tending to reduce the drift time, but it also increases the penetration of the electrons into the drift space, tending to increase drift time. For a linear potential variation in the drift space, the drift time decreases with increase in resonator voltage and the frequency increases with resonator voltage when the cathode-to-resonator voltage is numerically greater than the cathode-to-repeller voltage, while the drift time increases and the frequency decreases with increase in resonator voltage when the cathode-to-repeller voltage is numerically greater than the cathode-to-resonator voltage.[9] Changes in resonator voltage change the power input into the tube, and hence change the heating. This may change the frequency by changing the resonator size or shape. Hence, change in frequency may be different for slow changes in resonator voltage than for rapid changes in resonator voltage.

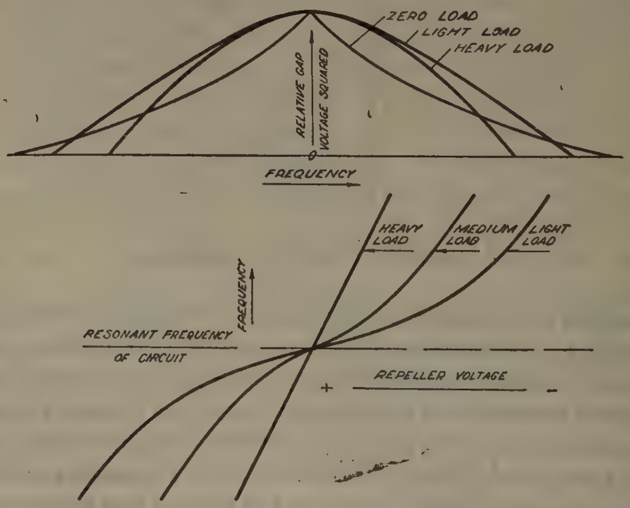


Fig. 8—Electronic-tuning curves.

VI. Modulation Coefficient

Imagine an electron crossing a gap across which a radio-frequency voltage of magnitude V appears. If the electron takes an appreciable part of a cycle to cross the gap, it will gain an energy less than V electron volts, regardless of the moment at which it crosses the gap. The peak energy it can gain may be expressed

[9] See Appendix B.

as MV, and M is called the modulation coefficient of the gap.

In crossing the gap the electron transfers a charge q from one side of the circuit to the other. This charge q times the voltage V is the work the electron does on the circuit. The work done on the electron is MVe, where e is the electronic charge. By the conservation of energy we see that $q = Me$.

The modulation coefficient, then, is a factor which relates the energy (measured in electron volts) gained by an electron crossing a gap to the voltage appearing across the gap, and also relates the circuit current induced in the circuit attached to the gap (of which the gap capacitance forms a part) to the electron current crossing the gap. We see that the maximum value of M is unity, for very short gaps, and it can be shown that M gets progressively smaller and may oscillate in sign as the transit time across the gap is increased. The exact variation of M with transit time depends on the configuration of the gap.

It may be seen that in the operation of a reflex oscillator the factor M is encountered twice. For a given gap voltage the amount of velocity modulation, and hence the strength of the radio-frequency electron current due to bunching are proportional to M. Further, the circuit current produced by the bunched electron current is proportional to M. Thus the circuit current for a given gap voltage is doubly dependent on M, and for small signals is proportional to M^2. The falling off in M because of increased gap transit time is partly responsible for the rapid decrease in power and in electronic-tunning range as resonator voltage is lowered.

VII. Influence of Load

It is well known that frequency of oscillation may be influenced by the load coupled to an oscillator. One kind of influence is obvious. A reactive load coupled to the resonant circuit changes the resonant frequency of that circuit, and hence the frequency of oscillation. Another kind of frequency change with load is particularly important in the case of reflex oscillators which have a wide electronic-tuning range. This can be explained by means of the frequency-versus-repeller-voltage diagrams of Fig. 8. Imagine, for instance, that the oscillator is operating off circuit resonance by means of electronic tuning. Such operation is represented by a point away from the common intersection of the frequency-versus-repeller-voltage curves of Fig. 8. Thus it may be seen that if the repeller voltage is kept constant and if the circuit conductance is changed, that is, the load is changed in purely resistive sense, without changing the resonant frequency of the resonant circuit, the frequency of oscillation will change in shifting to a new frequency-versus-repeller-voltage curve.

Not only frequency of operation is affected by load; the electronic tuning is affected as well. For instance, coupling the oscillator tightly to a high-Q resonant circuit has the effect of increasing the effective capacitance of the resonant circuit as defined in terms of the rate of change of gap susceptance versus frequency in the neighborhood of resonance or in terms of stored energy for a given radio-frequency gap voltage. This cuts down the electronic-tuning range and makes the variation of frequency with repeller voltage less rapid, as explained in Section IV.

Appendix A

Power Production

While the production of power in a reflex oscillator can be treated from the point of view of velocity modulation and bunching, as exemplified by the work of Webster,[4] it is interesting to note that a method of analysis of high-frequency tube behavior considerably antedating these ideas may be used in obtaining the desired result without any explicit reference to the concepts of velocity modulation, drift, or bunching. An excellent idea of early methods of analysis may be obtained from a paper by Benham.[10]

Assume that before crossing the gap of a reflex-type tube an electron has a velocity specified by a potential V_0. Assume that there is a radio-frequency voltage $V \sin \omega t_0$ across the gap, and that an electron crossing the gap gains an energy $MV \sin \omega t_0$ volts. Assume that after crossing the gap the electron enters a uniform field of strength E_0. The electron will then return to the gap in a time τ.

$$\tau = \left[2\sqrt{2(e/m)(V_0 + MV \sin \omega t_0)}\right]/(e/m)E_0 \qquad (1)$$

If $(V/V_0) \ll 1$, this may be written

$$\tau \doteq \tau_0(1 + (MV \sin \omega t_0/2V_0)). \qquad (2)$$

Here τ_0 is the drift time in absence of radio-frequency voltage. In a time dt_0 at t_0, a charge $-I_0 dt_0$ will cross the gap, I_0 being the direct beam current. This charge will pass the gap again at a time $t_0 + \tau$. The total element of work dW done in the two transits may be expressed

$$dW = (-I_0 dt_0)(MV)[\sin \omega t_0 - \sin \omega(t_0 + \tau_0(1 + (MV \sin \omega t_0/2V_0)))]. \qquad (3)$$

This may be written in terms of convenient quantities

$$\omega\tau_0 = \theta$$
$$\omega t_0 = \gamma$$

$$dW = -(I_0MV)/\omega[\sin \gamma - \sin(\gamma + \theta + (MV\theta \sin \gamma/2V_0))]d\gamma$$
$$(-I_0MV)/\omega(\sin \gamma - \sin \theta[\cos(\gamma + (MV\theta/2V_0) \sin \gamma)] - \cos \theta[\sin(\gamma + (MV\theta/2V_0) \sin \gamma)]. \qquad (4)$$

The work per cycle may be obtained by integrating dW from $\gamma = 0$ to $\gamma = 2\pi$. This integral times f, the frequency, is the power absorbed by the electron stream

$$P = f\int_0^{2\pi} dW = I_0MV \sin \theta J_1(MV\theta/2V_0). \qquad (5)$$

From this result it is easy to infer an electronic admittance

$$Y_e = (2I_0M/V)J_1(MV\theta/2V_0)e^{-j(\theta-\pi/2)}. \qquad (6)$$

[10] W. E. Benham, "A contribution to tube and amplifier theory," Proc. I.R.E., vol. 26, pp. 1093–1171; September, 1938.

Appendix B

Variation of Drift Time with Resonator Voltage

Suppose an electron enters a uniform field through a grid at a potential V_0 with respect to the cathode and travels toward a repeller a distance L away which has a potential $-V_R$ with respect to the cathode. The electron will penetrate a distance

$$x = LV_0/(V_0 + V_R). \qquad (7)$$

The time required to travel this distance and back will be

$$\tau_0 = 4x/\sqrt{2(e/m)V_0}$$
$$= (4L/\sqrt{2(e/m)})\sqrt{V_0}/(V_0 + V_R). \qquad (8)$$

Differentiating with respect to V_0 and combining the expression so obtained with (8) we obtain

$$d\tau_0/dV_0 = (\tau_0/2V_0)(V_0 - V_R)/(V_0 + V_R). \qquad (9)$$

From time to time it is proposed to present in the PROCEEDINGS papers of tutorial nature which, from the clarity of their presentation, the comprehensiveness of their data, and the convenience of their availability in the PROCEEDINGS are suitable for that purpose. Accordingly the following paper has been placed before the Institute membership.

The Editor

The Theory of Transmission Lines*

EDWARD N. DINGLEY, JR.†, SENIOR MEMBER, I.R.E.

***Summary*—Most texts dealing with this subject assume that the reader is familiar with complex exponential functions. They also omit steps in mathematical transformations on the assumption that the reader can interpolate these steps. The purpose of this article is not to present new material but only to describe, in step-by-step fashion, the derivation of the formulas necessary to the solution of transmission-line problems and to demonstrate, by examples, how to use these formulas. This treatment will be helpful to students and to engineers not regularly confronted with transmission-line problems.**

In an infinitely long uniform transmission line having uniformly distributed constants of $X=\omega L$ ohms series reactance, R ohms series resistance, G ohms leakage conductance, and $B=\omega C$ ohms shunt susceptance per unit loop length, the series impedance may

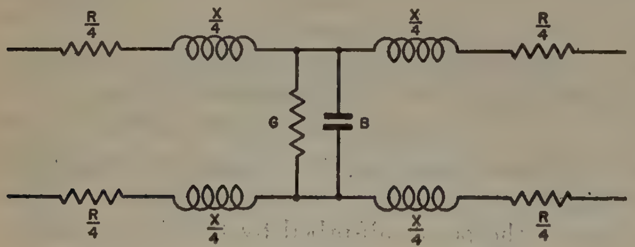


Fig. 1

be termed $Z=R+jX$ and the shunt acceptance may be termed $Y=G+jB$ per unit loop length. In the foregoing $\omega=2\pi f$, f=frequency in cycles per second, L=inductance in henries, and C=capacitance in farads, per unit loop length of line. A unit loop length is a unit length of the transmission line the series resistance of which is the sum of the series resistances of the two wires, the series reactance of which is the sum of the series reactances of each wire minus their mutual reactance, the shunt leakage conductance of which is the reciprocal of the leakage resistance between the two wires, and the shunt susceptance of which is the reciprocal of the capacitive reactance between the two wires. A unit length of line is shown in Fig. 1.

In an infinitesimally short length (dl) of the transmission line the voltage drop (dE) along the line (in the direction of the current flow) due to the series impedance will be the product of the current flowing (I) times the impedance of the line per unit length (Z) times the length of the line (dl) through which the current flows. In the same manner, the current lost along the line (in the direction of the current flow) due to the shunt acceptance will be the product of the voltage (E) acting across the line times the shunt acceptance (Y) per unit length times the length of the line (dl). By convention, voltage drop and current losses carry negative signs; therefore the above statements may be expressed algebraically as follows:

$$dE = -IZ(dl) \quad (1) \qquad dI = -EY(dl) \quad (1a)$$

or

$$dE/(dl) = -IZ \quad (2) \qquad dI/(dl) = -EY. \quad (2a)$$

Equations (2) and (2a) may be differentiated to obtain

$$d^2E/(dl)^2 = -Z(dI/(dl)) \qquad (3)$$
$$d^2I/(dl)^2 = -Y(dE/(dl)). \qquad (3a)$$

* Decimal classification: R116. Original manuscript received by the Institute, October 9, 1944.

† Bureau of Ships, Navy Department, Washington 25, D. C.

Equation (2) may be substituted in (3a) and (2a) may be substituted in (3), to obtain

$$d^2E/(dl)^2 = ZEY \quad (4) \qquad d^2I/(dl)^2 = ZIY. \quad (4a)$$

Equations (4) and (4a) are simple linear differential equations of the second order, the general solution for which is of the form

$$E = A_1e^{l\sqrt{ZY}} + B_1e^{-l\sqrt{ZY}} \quad (5)$$

$$I = A_2e^{l\sqrt{ZY}} + B_2e^{-l\sqrt{ZY}} \quad (5a)$$

where e = the base of the Napierian logarithms = 2.7183 –.

If (5) and (5a) are true, they must be true for all lengths of the line including a zero length. To find the constants of integration A_1, B_1, A_2, and B_2, place $l=o$ in (5) and (5a) and noting that $e^o=e^{-o}=1$ there is obtained:

$$E_s = A_1 + B_1 \quad (6) \qquad I_s = A_2 + B_2 \quad (6a)$$

where the subscript (s) refers to the sending end of the line because $l=o$.

Equations (5) and (5a) may be differentiated to obtain

$$dE/(dl) = A_1\sqrt{ZY}\, e^{l\sqrt{ZY}} - B_1\sqrt{ZY}\, e^{-l\sqrt{ZY}} \quad (7)$$

$$dI/(dl) = A_2\sqrt{ZY}\, e^{l\sqrt{ZY}} - B_2\sqrt{ZY}\, e^{-l\sqrt{ZY}}. \quad (7a)$$

In (7) and (7a), if $l=o$ then

$$dE_s/(dl) = A_1\sqrt{ZY} - B_1\sqrt{ZY} \quad (8)$$

$$dI_s/(dl) = A_2\sqrt{ZY} - B_2\sqrt{ZY}. \quad (8a)$$

Equation (2) may be substituted in (8) and (2a) may be substituted in (8a) to obtain

$$-I_sZ = \sqrt{ZY}\,(A_1 - B_1) \quad (9)$$

$$-E_sY = \sqrt{ZY}\,(A_2 - B_2) \quad (9a)$$

or by rearranging

$$-I_s\sqrt{Z/Y} = A_1 - B_1 \quad (10)$$

$$-E_s\sqrt{Y/Z} = A_2 - B_2. \quad (10a)$$

Equations (6) and (10) may be added and (6a) and (10a) may be added to obtain

$$A_1 = 1/2[E_s - I_s\sqrt{Z/Y}] \quad (11)$$

$$A_2 = 1/2[I_s - E_s\sqrt{Y/Z}]. \quad (11a)$$

Equations (6) and (10) may be subtracted and (6a) and (10a) may be subtracted to obtain

$$B_1 = 1/2[E_s + I_s\sqrt{Z/Y}] \quad (12)$$

$$B_2 = 1/2[I_s + E_s\sqrt{Y/Z}]. \quad (12a)$$

Equations (11) and (12) may be substituted in (5), and (11a) and (12a) may be substituted in (5a), to obtain

$$E = 1/2[E_s - I_s\sqrt{Z/Y}]e^{l\sqrt{ZY}} + 1/2[E_s + I_s\sqrt{Z/Y}]e^{-l\sqrt{ZY}} \quad (13)$$

$$I = 1/2[I_s - E_s\sqrt{Y/Z}]e^{l\sqrt{ZY}} + 1/2[I_s + E_s\sqrt{Y/Z}]e^{-l\sqrt{ZY}}. \quad (13a)$$

If the line is infinitely long ($l=\infty$) then E and I at the far end must be zero. Note that $e^{-\infty}=o$. Then there is obtained from (13) and (13a)

$$o = E_se^{\infty} - I_s\sqrt{Z/Y}\,e^{\infty} \quad (14)$$

$$o = I_se^{\infty} - E_s\sqrt{Y/Z}\,e^{\infty}. \quad (14a)$$

Rearranging (14) and (14a) there is obtained

$$E_s/I_s = \sqrt{Z/Y} \quad (15) \qquad E_s/I_s = \sqrt{Z/Y}. \quad (15a)$$

But E_s/I_s is the ratio of sending voltage to sending current at the input of an infinitely long line and is therefore the impedance of an infinitely long line. This impedance ($\sqrt{Z/Y}$) is called the "surge" impedance and is written "Z_o."

Equations (13) and (13a) state that if the voltage (E_s) and the current (I_s) at any point (called the sending point) on an infinitely long line are known, the voltage and current at another point in the direction of current flow may be determined. Call this second point R (the receiving point) and call the voltage and current E_R and I_R. If the voltage and current at a point in the direction opposed to the current flow is desired, then the distance (l) to that point must be written ($-l$) and the positions of the sending and receiving points become transposed. Thus, in (13) and (13a), if $-l$ is substituted for $+l$ *and* E_R and I_R are substituted for E_s and I_s, respectively, there is obtained

$$E=1/2[E_R+I_RZ_o]e^{l\sqrt{ZY}}+1/2[E_R-I_RZ_o]e^{-l\sqrt{ZY}} \quad (16)$$

$$I=1/2[I_R+E_R/Z_o]e^{l\sqrt{ZY}}+1/2[I_R-E_R/Z_o]e^{-l\sqrt{ZY}}. \quad (16a)$$

In (16) and (16a) assume that the voltage E_R and the current I_R exist at the point R on a line which is infinitely long to the right of this point and which is finitely long to the left of this point. As the right-hand portion is infinitely long, it represents a constant impedance Z_o in shunt to the line at the point R but which has no effect on the line to the left of this point because E_R and I_R have been stated to exist regardless of impedances which may be shunted across the point R. Thus the infinitely long right-hand portion of the line may be considered as nonexistent, and the finite left-hand portion of the line may be considered as being terminated at the receiving point R and as being terminated in the impedance Z_R such that $Z_R=E_R/I_R$.

Equation (16) may be multiplied by $2/I_R$ and (16a) may be multiplied by $2Z_R/E_R$ to obtain

$$2E/I_R= [E_R/I_R+Z_o]e^{l\sqrt{ZY}}+[E_R/I_R-Z_o]e^{-l\sqrt{ZY}} \quad (17)$$

$$2IZ_R/E_R= [I_RZ_R/E_R+Z_R/Z_o]e^{l\sqrt{ZY}} + [I_RZ_R/E_R-Z_R/Z_o]e^{-l\sqrt{ZY}}. \quad (17a)$$

But, $E_R/I_R=Z_R$ and $I_R/E_R=1/Z_R$, therefore,

$$2E/I_R= [Z_R+Z_o]e^{l\sqrt{ZY}}+[Z_R-Z_o]e^{-l\sqrt{ZY}} \quad (18)$$

$$2IZ_R/E_R= [1+Z_R/Z_o]e^{l\sqrt{ZY}}+[1-Z_R/Z_o]e^{-l\sqrt{ZY}}. \quad (18a)$$

Equation (18) may be divided by (18a) to obtain

$$\frac{E}{I} = Z_s = Z_o\frac{[Z_R + Z_o]e^{l\sqrt{ZY}} + [Z_R - Z_o]e^{-l\sqrt{ZY}}}{[Z_R + Z_o]e^{l\sqrt{ZY}} - [Z_R - Z_o]e^{-l\sqrt{ZY}}}. \quad (19)$$

Equation (19) is the exact expression for the input impedance of a transmission line having the length l, the series impedance Z per unit length, the shunt admittance Y per unit length, and the surge impedance Z_o, which is terminated in the impedance Z_R.

In order for (19) to be useful it is necessary to define the terms $e^{l\sqrt{ZY}}$ and $e^{-l\sqrt{ZY}}$.

The first paragraph of this article defined $Z=R+jX$ and $Y=G+jB$. It follows that

$$\sqrt{ZY}=\sqrt{(R+jX)(G+jB)}. \tag{20}$$

The right-hand member of (20) will contain real and quadrature components. If these components are called α and $j\beta$, respectively, (20) becomes

$$\alpha+j\beta=\sqrt{(R+jX)(G+jB)}=\sqrt{ZY} \tag{21}$$

$$(\alpha+j\beta)^2=(R+jX)(G+jB) \tag{22}$$

$$\alpha^2-\beta^2+j2\alpha\beta=(RG-XB)+j(GX+BR). \tag{23}$$

(Note: $j\times j=-1$).

The real and quadrature terms may be equated to obtain

$$\alpha^2-\beta^2=(RG-XB) \tag{24}$$

$$2\alpha\beta=(GX+BR) \tag{24a}$$

or, from (24a),

$$\alpha=(GX+BR)/2\beta. \tag{24b}$$

Equation (24b) may be substituted in (24) to obtain

$$[(GX+BR)^2]/4\beta^2-\beta^2=(RG-XB) \tag{25}$$

or, simplifying

$$\beta^4+\beta^2(RG-XB)-(GX+BR)^2/4=o. \tag{26}$$

Solving for β and α

$$\beta=\sqrt{1/2[\sqrt{[(R^2+X^2)(G^2+B^2)]^{\frac{1}{2}}}-(RG-XB)]} \quad \text{circular radians} \tag{27}$$

$$\alpha=\sqrt{1/2[\sqrt{[(R^2+X^2)(G^2+B^2)]^{\frac{1}{2}}}+(RG-XB)} \quad \text{hyperbolic radians.} \tag{28}$$

If the values of R, X, G, and B per unit length are known, then α and β may be computed by (27) and (28) and Z_o may be computed from (15) as follows:

$$Z_o=\sqrt{Z/Y}=\sqrt{(R+jX)/(G+jB)}. \tag{29}$$

In (27) and (28), β is an angle measured in circular radians and, as will be demonstrated later, represents the angular change in phase of the voltage or current as it travels a unit length of the transmission line and α, measured in hyperbolic radians, represents the change in amplitude of the voltage or current as it travels a unit length of the transmission line.

If $l\beta$ represents the difference in phase of the voltage (or current) at the point l compared to the voltage (or current) at the point o, then the voltage (or current) may be considered as being a sine wave of energy traveling down the line and the crest of a wave leaving o will reach l in the length of time (t) required for the rotating vector representing the voltage (or current) to rotate through $l\beta$ radians. This vector, however, is rotating at the rate $\omega=2\pi f$ radians per second; therefore $t=l\beta/2\pi f$ seconds. If the crest of the wave has traveled the distance l in $t=l\beta/2\pi f$ seconds, then the velocity with which the wave travels on the wire is

$$V=l/t=2\pi f/\beta=\omega/\beta \text{ unit lengths per second.} \tag{30}$$

It will be noted that the velocity of travel of the wave on the line is proportional to the frequency f and inversely proportional to β. If, as is usually the case in telephone lines, β is not also proportional to f, then waves of different frequencies will travel at different velocities along the line and will reach the receiving end of the line with phase relations not identical to those at the sending end, and serious distortion will result.

The term β is called the "wavelength constant" of the transmission line for the reason that if the phase of the voltage (or current) along the line changes β radians per unit length, then the number of unit lengths required for the phase to shift 2π radians will be

$$\lambda=2\pi/\beta. \tag{31}$$

This number of unit lengths is called λ because it is the distance between two points on the transmission line having the same phase or the same instantaneous values of voltage or current. If β is in radians per loop meter of line length, then λ will be in meters.

The term α is called the "attenuation constant" of the transmission line for the reason that it represents the change in the crest amplitude or the change in the effective value of the voltage or current as it travels a unit length of the transmission line.

The vector sum of α and β, expressed as $\alpha+j\beta$ is called the "propagation constant" of the line. These terms are not "constants" because they vary with frequency. In (27) and (28), which express the exact values of α and β, it will be noted that both $X=\omega L$ and $B=\omega C$ are discrete functions of frequency while R and G both vary somewhat with frequency because of "skin effect" and variable dielectric losses which are functions of frequency.

There are two special cases relating to α and β which are of particular interest. The most important case concerns relatively short radio-frequency transmission lines wherein, to a close approximation, R and G are negligibly small compared to X and B. If $R=o=G$ is substituted in (27), (28), and (29), there is obtained

$$\beta=\sqrt{XB}=\sqrt{(\omega L)(\omega C)}=\omega\sqrt{LC} \tag{32}$$

$$\alpha=0 \tag{33}$$

$$Z_o=\sqrt{jX/jB}=\sqrt{\omega L/\omega C}=\sqrt{L/C} \tag{34}$$

where $\omega=2\pi f$,
f = frequency in cycles per second
L = inductance of line in henries per unit loop length
C = capacitance of line in farads per unit loop length
$X=\omega L$
$B=\omega C$.

It has been demonstrated[1] that the inductance in henries and capacitance in farads per centimeter length of two parallel wires is[2]

[1] G. W. Pierce, "Electric Oscillations and Electric Waves," first edition; McGraw-Hill Book Company, New York 18, N. Y., pp. 332–333.

[2] Equations (35) and (36) are accurate only when d is at least 20 times greater than r. For a coaxial line, if the inner radius of the outer conductor is substituted for d and the outer radius of the inner conductor is substituted for r, then the inductance of the coaxial line per centimeter of length will be one half of equation (35) and the capacitance will be twice equation (36) and the formulas are exact for any values of d and r.

$$L=\frac{4\mu \log_e (d/r)}{10^9} \text{ henries per centimeter loop length} \quad (35)$$

$$C=\frac{K10^9}{4(3\times10^{10})^2 \log_e (d/r)} \text{ farads per centimeter loop length.} \quad (36)$$

where $\log_e$ = Napierian logarithm

d = axial distance between wires in any units

r = radius of each wire in same units

μ = magnetic permeability of medium between wires

K = dielectric constant of medium between wires

3×10^{10} = ratio of electromagnetic units to electrostatic units = centimeters per second.

In the case of the 2-wire line, if the medium between the wires is air, then $K=1=\mu$, and from (35) and (36)

$$\left.\begin{aligned}\sqrt{LC} &= 1/(3\times10^{10}) \quad (37)\\ \sqrt{L/C} &= 120 \log_e (d/r) \quad (38)\end{aligned}\right\} \text{2-wire lines.}$$

In the case of the coaxial line, if the medium between the wires is air, then $K=1=\mu$, and from (35) and (36) and the footnote 2

$$\left.\begin{aligned}\sqrt{LC} &= 1/(3\times10^{10}) \quad (39)\\ \sqrt{L/C} &= 60 \log_e (d/r) \quad (40)\end{aligned}\right\} \text{coaxial lines.}$$

Substituting (37) and (39) in (32) and (38) and (40) in (34) there are obtained the following values for *this special case* of $R=o=G$:

$$\beta = \omega/(3\times10^{10}) \quad (41)$$

$$\alpha = o \quad (42)$$

$$Z_o(\text{2-wire}) = 120 \log_e (d/r) \text{ ohms} \quad (43)$$

$$Z_o(\text{coaxial}) = 60 \log_e (d/r) \text{ ohms.} \quad (44)$$

Substituting (41) in (30) and (31) there is obtained

$$V = \frac{\omega}{\omega/(3\times10^{10})} = 3\times10^{10} \text{ centimeters per second} \quad (45)$$

$$\lambda = \frac{2\pi}{\omega/(3\times10^{10})} = \frac{2\pi(3\times10^{10})}{2\pi f} = \frac{3\times10^{10}}{f} \text{ centimeters.} \quad (46)$$

Thus, in a line having $R=o=G$, the velocity of propagation is 3×10^{10} centimeters per second and is independent of frequency, the same as in air; the wavelength, or distance between crests of the waves, equals $3\times10^{10}/f$ centimeters (f in cycles) or equals $3\times10^5/f$ meters (f in kilocycles), the same as in air; and the attenuation of the voltage and current is zero, that is, the sending and receiving voltages (and currents) are equal. These values of V and λ are valid whether or not there are "standing waves" on the line. Standing waves will be defined later.

On telephone lines using voice frequencies it is desirable that the velocity of propagation should be the same for all frequencies. This condition can be attained if $R=o=G$ as demonstrated above, but such conditions are physically impossible. The alternative leads to the second important case involving α and β.

In most telephone lines and cables, R and B are much larger than G and X; that is, R and C are much larger than G and L. It is desirable that $R\times B$ should equal $G\times X$. This is partially accomplished by reducing R and B by using larger diameter wires spaced farther apart. It is obviously undesirable to increase the leakage conductance G to attain the desired equality but the inductance L can be increased by loading the line at intervals with lumped inductors, as is done with long land lines or by uniformly loading the line by wrapping it with high-permeability material, as is done with cables. If $R\times B$ is made to equal $G\times X$, and these values are substituted into (27) and (28), there is obtained

$$\beta = \sqrt{XB} = \omega\sqrt{LC} \quad (47)$$

$$\alpha = \sqrt{GR}. \quad (48)$$

Equation (47) is identical to (32) and therefore the velocity of propagation in this special case of $R\times B=G\times X$ is the same as in the special case of $R=o=G$. This velocity is constant for all frequencies, is the same as the velocity of transmission in air, and is correctly expressed by (45). Similarly, the wavelength or distance between wave crests is the same in this case as in the other and is the same as the wavelength of transmissions in air and is correctly expressed by (46).

Equation (48) is not equal to zero as was (42) of the first case and therefore represents a finite attenuation.

Having defined the term $\sqrt{YZ}$, the propagation constant of the line, as being equal to $\alpha+j\beta$ and having further defined α and β in terms of R, G, X, and B of the line, and noting that $e^{l\sqrt{ZY}}=e^{l(\alpha+j\beta)}=e^{\alpha l}e^{j\beta l}$, equations (16) and (16a) may be rewritten as follows:

$$E=1/2[E_R+I_RZ_o]e^{\alpha l}e^{j\beta l}+1/2[E_R-I_RZ_o]e^{-\alpha l}e^{-j\beta l} \quad (49)$$

$$I=1/2[I_R+E_R/Z_o]e^{\alpha l}e^{j\beta l}+1/2[I_R-E_R/Z_o]e^{-\alpha l}e^{-j\beta l}. \quad (50)$$

In analyzing the meaning of (49) and (50) it should be noted that I_RZ_o is a vector voltage which, when added to the vector voltage E_R represents a vector voltage (say E_1) making a definite phase angle (say ϕ_1) with respect to a reference axis. In the same manner $E_R-I_RZ_o$ is another vector voltage (say E_2) making a definite phase angle (say ϕ_2) with the same reference axis. In vector notation, the first vector is represented by $E_1e^{j\phi_1}$ and the second by $E_2e^{j\phi_2}$ where the operator $e^{j\phi}$ is merely a shorthand method of stating that the vector to which it is attached has been rotated counterclockwise (leading phase) by an angle ϕ. The expression $e^{-j\phi}$ means that the vector is rotated clockwise (lagging phase). The expression $Ee^{l\alpha}$ is merely a shorthand way of stating that the vector of length E has been lengthened by multiplying it by $e^{l\alpha}$ or by 2.7183 raised to the $l\alpha$ power. As $e^{-l\alpha}=1/e^{l\alpha}$ is less than unity, this operator shortens the vector.

Equation (49) can now be written as follows:

$$E = 1/2E_1e^{\alpha l}e^{j\beta l}e^{j\phi_1} + 1/2E_2e^{-\alpha l}e^{-j\beta l}e^{j\phi_2}. \quad (51)$$

Of the right-hand member of (51), the first term represents a vector voltage which has an initial amplitude $E_1/2$ and initial phase ϕ_1, when $l=o$, and which increases in amplitude by $e^{\alpha l}$ and advances in phase by the angle $e^{j\beta l}$ as l increases. Because l increases from the

receiving end toward the transmitting end of the line, this right-hand member represents a voltage wave which attenuates and lags in phase as it travels from *sending* to *receiving* end of the line and is therefore the sent wave. In the same manner, the second term of (51) represents a voltage which attenuates and lags in phase as it travels from the *receiving* end to the *sending* end and therefore constitutes a wave reflected from the receiving end.

These transmitted and reflected waves may be represented by the vectors in Fig. 2. At the point $l=o$, the transmitted wave $E_1/2$ is represented by a vector (A) of *constant length* rotating counterclockwise about the point o at the angular velocity of $\omega=2\pi f$ radians per second. Its projection at any instant on the $X-X$ axis represents the instantaneous voltage at the point $l=0$

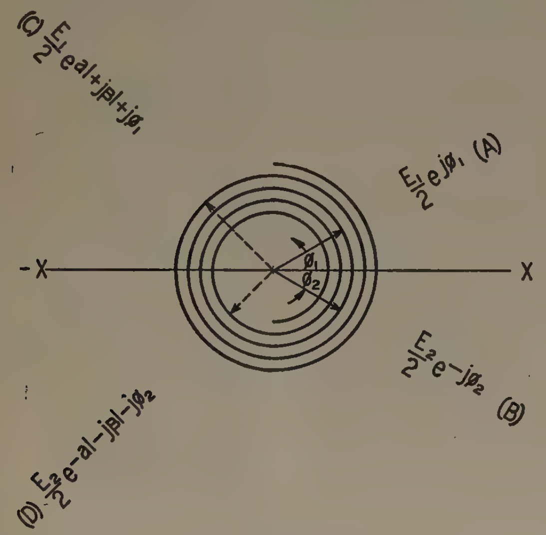


Fig. 2

on the line due to the transmitted wave. As the vector (A) rotates at constant angular speed, its projection on $X-X$ axis represents a voltage which alternates sinusoidally.

In the same manner the reflected wave $E_2/2$ is represented by the vector (B) of *constant length* also rotating counterclockwise about the point o at the same angular velocity $\omega=2\pi f$ radians per second. Its projection at any instant on the $X-X$ axis represents the instantaneous voltage at the point $l=0$ on the line due to the *reflected wave*. This projection also represents a voltage which alternates sinusoidally. The total instantaneous voltage on the line at the point $l=0$ is the sum of the projections on $X-X$ of vectors (A) and (B) taken at that instant. By inspection it is seen that the sum of these two projections would be a maximum and equal to $E_1/2+E_2/2$ if $\phi_1+\phi_2=0$, that is, if the vectors (A) and (B) always coincided as they rotated, and it would be a minimum and equal to $(E_1/2)-(E_2/2)$ if $\phi_1+\phi_2=\pi=180$ degrees, that is, if the vectors were always opposite each other as they rotated.

If a point $l=l$ on the line is considered, then the vector (A) has gained in phase and the vector (B) has lagged in phase until their relative positions are as shown at (C) and (D) in Fig. 2. However, in advancing phase, the vector (A) must follow the *spiral* in a counterclockwise direction which increases the amplitude from $E_1/2$ to $[E_1/2e]^{\alpha l}$ and in retarding phase, the vector (B) must follow the spiral in a clockwise direction which decreases its amplitude from $E_2/2$ to $[E_2/2]e^{-\alpha l}$. In Fig. 2 the vectors (A) and (B) have been rotated through approximately $5\pi/2$ radians or 450 degrees in opposite directions in order to reach the positions shown by vectors (C) and (D).

If our movement along the line from $l=o$ to $l=l$ is stopped at the point $l=l$, then the two vectors cease to gain and lose phase and length with respect to each other and may be considered to be two new vectors both of which rotate counterclockwise at the rate $\omega=2\pi f$ and the crest or maximum voltage obtained at the point $l=l$ will be the maximum of the sum of the projections of these vectors on the axis $X-X$.

Thus it has been demonstrated that there is an interference pattern produced on the line by the transmitted wave and reflected wave; that at certain points on the line the crest of the resultant wave is equal to the sum of the two and at other points it is equal to their difference. The greatest maxima and minima will occur near the receiving end where the transmitted wave and received wave are nearly equal in amplitude.

The foregoing analysis applies to the current waves in exactly the same manner.

The following special cases involving (49) and (50) are of particular interest:

Infinitely Long Line. Here $l=\infty$, therefore $e^{-\alpha l}=e^{-\infty}=o$, therefore the second terms of the right-hand members of (49) and (50) are zero and therefore there are no reflected waves. In this case there are no consecutive points of maximum and minimum voltage or current along the line; on the contrary, the voltage and current increase smoothly from the received value to infinity at the point $l=\infty$.

Open-Circuited Line. Here $I_R=o$ because the receiving end of the line is open. Equations (49) and (50) then reduce to

$$E = E_R\left[\frac{e^{\alpha l+j\beta l}+e^{-\alpha l-j\beta l}}{2}\right] \tag{49a}$$

$$I = \frac{E_R}{Z_o}\left[\frac{e^{\alpha l+j\beta l}-e^{-\alpha l-j\beta l}}{2}\right]. \tag{50a}$$

If $\beta l=\pi/2$, $3\pi/2$, $5\pi/2$ etc., then,

$$E = E_R\left[\frac{e^{\alpha l}-e^{-\alpha l}}{2}\right] = 0 \quad \text{if} \quad \alpha = 0 \tag{49b}$$

$$I = j\frac{E_R}{Z_o}\left[\frac{e^{\alpha l}+e^{-\alpha l}}{2}\right] = j\frac{E_R}{Z_o} \quad \text{if} \quad \alpha = 0. \tag{50b}$$

If, from the above, $l=\pi/2\beta$, $3\pi/2\beta$, $5\pi/2\beta$, etc., and if $\alpha=0$, equation (31) may be substituted for β above and there is obtained $l=[1/4]\lambda$, $[3/4]\lambda$, $[5/4]\lambda$, etc.

Thus for an open-circuited line, voltage minima occur at a point one-quarter wavelength from the open end and at every half wavelength thereafter. Current maxima

occur at these same points. The voltage and current are in quadrature as indicated by the j preceding the current term. Note here that $E_R e^{j(\pi/2)}$ is a polar co-ordinate shorthand method of designating a vector rotated 90 degrees $=\pi/2$ radians from the reference axis. The term jE_R designates the same thing in rectangular co-ordinate form. The two, therefore, are equal.

Short-Circuited Line. Here $E_R=0$ because the receiving end of the line is short-circuited. Equations (49) and (50) then reduce to

$$E = I_R Z\left[\frac{e^{\alpha l+j\beta l} - e^{-\alpha l-j\beta l}}{2}\right] \tag{49c}$$

$$I = I_R\left[\frac{e^{\alpha l+j\beta l} + e^{-\alpha l-j\beta l}}{2}\right]. \tag{50c}$$

If $\beta l=\pi/2$, $3\pi/2$, $5\pi/2$ etc., then,

$$E = jI_R Z_o\left[\frac{e^{\alpha l} + e^{-\alpha l}}{2}\right] = jI_R Z_o \quad \text{if} \quad \alpha = 0 \tag{49d}$$

$$I = I_R\left[\frac{e^{\alpha l} - e^{-\alpha l}}{2}\right] = 0 \quad \text{if} \quad \alpha = 0. \tag{50d}$$

If, from the above, $l=\pi/2\beta$, $3\pi/2\beta$, $5\pi/2\beta$, etc., and if $\alpha=0$, equation (31) may be substituted for β above and there is obtained: $l=[1/4]\lambda$, $[3/4]\lambda$, $[5/4]\lambda$, etc.

Thus, for a short-circuited line, voltage maxima occur at a point one-quarter wavelength from the short-circuited end and at every half wavelength thereafter. Current minima occur at these same points. The voltage and current are in quadrature as indicated by the j preceding the voltage term.

It will be noted that the distance between any two voltage (or current) nodes is equal to one half the wavelength of the waves on the line. If the velocity of propagation is equal to that in air as is usually the case with radio-frequency transmission lines, the frequency in kilocycles may be found by dividing 3×10^5 by the wavelength in meters.

Line Terminated in Z_o. Here the line is terminated in an impedance equal to Z_o, therefore the product of the received current (I_R) and the impedance through which it flows (Z_o) is equal to the received voltage E_R. In the same manner $E_R/Z_o=I_R$. Thus the second terms of the right-hand members of (49) and (50) become zero, there is no reflected wave, and the equations become

$$E = E_R e^{\alpha l} e^{j\beta l} = E_R e^{\alpha l+j\beta l} \tag{49e}$$

$$I = I_R e^{\alpha l} e^{j\beta l} = I_R e^{\alpha l+j\beta l}. \tag{50e}$$

Thus for a line terminated in an impedance equal in magnitude and phase to the surge impedance of the line, there are no reflected waves, and hence, no standing waves on the line, and the voltage and current at any point on the line have a phase difference equal to the phase angle of Z_o.

Line Input Impedance. Having defined the term $\sqrt{ZY}$, the propagation constant, as being equal to $\alpha+j\beta$ and having further defined α and β in terms of R, G, X, and B of the line, equation (19) may be written

$$Z_s = Z_o\frac{[Z_R + Z_o]e^{\alpha l+j\beta l} + [Z_R - Z_o]e^{-\alpha l-j\beta l}}{[Z_R + Z_o]e^{\alpha l+j\beta l} - [Z_R - Z_o]e^{-\alpha l-j\beta l}}. \tag{52}$$

If the values of the terms in (52) are known, the equation is easily solved as follows:

In general, Z_R and Z_o will each consist of a resistive (real) term and a reactive (quadrature) term. Add the reals and add or subtract the "quads" as required and obtain, say: $R+jQ$ in rectangular co-ordinates.

Convert to polar co-ordinates as follows:

$$R + jQ = \sqrt{R^2 + Q^2}\, e^{j\tan^{-1}(Q/R)} = \text{say,}\ Ze^{j\theta}$$

where $\tan^{-1} Q/R=\theta=$the angle whose tangent is Q/R.

Next convert βl from radians to degrees by multiplying by 57.3. Next add or subtract βl degrees to or from θ according to whether $j\beta l$ is plus or minus.

Next multiply or divide Z by the number $e^{\alpha l}$, depending on whether αl is plus or minus.

There is obtained a new vector making the angle $\theta\pm\beta l$ with the origin and having a length of $Ze^{\pm\alpha l}$. Call this new vector $Z_1e^{j\theta_1}$ in polar co-ordinates.

Convert to rectangular co-ordinates as follows:

$$Z_1e^{j\theta_1} = Z_1\cos\theta_1 + jZ_1\sin\theta_1 \quad \text{or say} \quad R_1 + jQ_1.$$

The four bracketed parts of (52) may be treated in this manner to convert each part to the form R_1+jQ_1.

Next add the real (R_1) terms of the numerator parts and add the quad (jQ) terms of the numerator parts to obtain new real and quad parts, say R_2+jQ_2. Do the same for the denominator.

Next convert the numerator to polar form, say $Z_2e^{j\theta_2}$ as demonstrated above. Do the same with the denominator.

Next arithmetically divide the numerator Z_2 term by the denominator Z_2 term and algebraically subtract their angles θ_2 remembering that $-(-\theta)=+\theta$ and obtain the form $Z_3{}^{j\theta_3}$.

Next place Z_o in its polar form ($Z_oe^{j\theta_o}$) as demonstrated above. Arithmetically multiply Z_o and Z_3 and algebraically add their phase angles θ_o and θ_3 to obtain the new form $Z_4e^{j\theta_4}$ which is the desired value of the input impedance of the line in *polar form*. This may be converted into the rectangular co-ordinate form of R_4+jQ_4 as demonstrated above. The latter form shows that Z_s will consist in general of an apparent resistance R_4 in series with an apparent reactance Q_4.

In (52) it should be noted that if the line is infinitely long ($l=\infty$) then $Z_s=Z_o$. Also if a line of finite length l is terminated in an impedance equal to Z_o then $Z_s=Z_o$.

Measurement of Z_o, α and β. If the values of R, G, X, and B are known, then Z_o, α and β can be determined from (27), (28), and (29). In most cases, the values of R, G, X, and B are not known and cannot be measured directly. The values of Z_o and α and β can, however, be determined by measuring Z_s (the apparent input impedance of the line) for the condition that the far end of the line is short-circuited and again for the condition that the far end of the line is open-circuited. Call these values Z_{sc} and Z_{oc}, respectively. The apparent impedance is measured in the same way that the apparent

impedance of an antenna is measured, that is, by driving the line through a reactance of such a value as to produce resonance, under which condition the apparent reactance of the line is equal to the tuning reactance but of the opposite sign or kind. At resonance, resistance is inserted until the input current is halved, under which condition the apparent input resistance of the line equals the inserted resistance. For a given uniform line, identical values of α and β (per unit length) and identical values of Z_o will be obtained regardless of the length of line which is measured. However, if the line has a length equal to any multiple of a quarter wavelength the measurement of Z_{oc} or Z_{sc} will be difficult because one will be nearly infinite while the other will be nearly zero. Consequently, the length of line or frequency of measurement should be chosen so that the length of the line is approximately an odd multiple of one-eighth wavelength.

In (52), if $Z_R=\infty$, then $Z_R+Z_o=Z_R$, and

$$Z_{oc}=Z_o\left[\frac{e^{\alpha l+j\beta l}+e^{-\alpha l-j\beta l}}{e^{\alpha l+j\beta l}-e^{-\alpha l-j\beta l}}\right] \tag{53}$$

and if $Z_R=0$ then

$$Z_{sc}=Z_o\left[\frac{e^{\alpha l+j\beta l}-e^{-\alpha l-j\beta l}}{e^{\alpha l+j\beta l}+e^{-\alpha l-j\beta l}}\right]. \tag{54}$$

If (53) and (54) are multiplied there is obtained

$$Z_o=\sqrt{Z_{oc}Z_{sc}}. \tag{55}$$

Equation (55) states that the surge impedance of any line is equal to the root of the vector product $Z_{oc}\times Z_{sc}$. To accomplish this operation, convert the measured values of Z_{oc} and Z_{sc} from the form $R+jX$ to the form $\sqrt{R^2+X^2}e^{j\tan^{-1}(X/R)}$ or say $Z_{oc}e^{j\phi_{oc}}$ and $Z_{sc}e^{j\phi_{sc}}$. The product of these terms is $Z_{oc}Z_{sc}e^{j(\phi_{oc}+\phi_{sc})}$ and the square root of this product is $\sqrt{Z_{oc}Z_{sc}}e^{j[(\phi_{oc}+\phi_{sc})/2]}=Z_o$. This polar expression for Z_o may be converted into the rectangular co-ordinate form of $R+jX$ by the method previously demonstrated.

If the root of the quotient of (54) divided by (53) is taken, there is obtained

$$\sqrt{\frac{Z_{sc}}{Z_{oc}}}=\frac{e^{\alpha l+j\beta l}-e^{-\alpha l-j\beta l}}{e^{\alpha l+j\beta l}+e^{-\alpha l-j\beta l}}. \tag{56}$$

The value of $\sqrt{Z_{sc}/Z_{oc}}$ is obtained from the measured values of Z_{sc} and Z_{oc} by using the same type of operations as were described for evaluating $\sqrt{Z_{sc}Z_{oc}}$.

Call the real and imaginary terms of $\sqrt{Z_{sc}/Z_{oc}}$, A and jM, respectively, and obtain

$$\sqrt{Z_{sc}/Z_{oc}}=A+jM. \tag{57}$$

It can then be demonstrated[3]

$$\tanh 2\alpha l=2A/(1+A^2+M^2) \tag{58}$$

and

$$\tan 2\beta l=2M/(1-(A^2+M^2)). \tag{59}$$

The value of αl may be found by taking one half of the hyperbolic angle whose hyperbolic tangent is (58) and the value of βl may be found by taking one half of the circular angle whose trigometric tangent is (59). Tables of trigonometric and hyperbolic tangents may be found in almost any electrical handbook. An hyperbolic angle is really a scaler quantity (a number) and may be treated as such. A circular angle is a real angle and is measured in electrical degrees or radians.

The following is a numerical example of the use of the foregoing formulas:

Assume a transmission line 23.3 meters long. Assume that the values of input impedance are measured at 1592 kilocycles and found to be

$$Z_{oc}=322-j560=629.5e^{j59°.2}\text{ ohms}$$
$$Z_{sc}=462+j427=629.5e^{j42°.8}\text{ ohms.}$$

The first expression means that the measured impedance Z_{oc} is composed of a resistance of 322 ohms in series with a capacitive reactance of 560 ohms. The second expression means that the measured impedance Z_{sc} is composed of a resistance of 462 ohms in series with an inductive reactance of 427 ohms. The alternative expressions are the polar forms of the same equations and state that Z_{oc} consists of an impedance of 629.5 ohms having a phase angle of $-59.°2$ that Z_{sc} consists of an impedance of 629.5 ohms having a phase angle of $+42.°8$.

The conversion of Z_{oc} from rectangular co-ordinate form to polar form is accomplished as follows:

$$322-j560=\sqrt{(322)^2+(560)^2}\,e^{j\phi}=629.5e^{j\phi}$$

where ϕ is the angle whose tangent is $-560/322=-59.2$ degrees.

The conversion of Z_{oc} from polar form to rectangular co-ordinate form is accomplished as follows:

$$629.5e^{-j59.°2}=629.5\cos 59.°2-j629.5\sin 59.°2$$
$$=322-j560.$$

Using (57)

$$\sqrt{\frac{Z_{sc}}{Z_{oc}}}=A+jM=\sqrt{\frac{629.5e^{j42.°8}}{629.5e^{-j59.°2}}}=\sqrt{1\times e^{j102°}}$$
$$A+jM=1\times e^{j51°}$$
$$=1\times[\cos 51°+j\sin 51°]$$
$$A+jM=0.629+j0.777$$

therefore, $A=0.629$ and $M=0.777$.

Using (58)

$$\tanh 2\alpha l=\frac{2A}{1+A^2+M^2}=\frac{2(0.629)}{1+(0.629)^2+(0.777)^2}=0.629$$

or $2\alpha l=0.74$ hyperbolic radians (from tables)
or $\alpha l=0.37$ hyperbolic radians.

Using (59)

$$\tan 2\beta l=\frac{2M}{1-(A^2+M^2)}=\frac{2(0.777)}{1-[(0.629)^2+(0.777)^2]}=\frac{1.554}{0}$$

or $2\beta l=\pi/2$ circular radians $=90$ degrees (from tables)
or $\beta l=\pi/4$ circular radians $=45$ degrees.

The length of the line is 2,330 centimeters, therefore

$$\alpha=\frac{0.37}{l}=\frac{0.37}{2330}$$
$$=1.587\times 10^{-4}\text{ hyperbolic radians per centimeter.}$$
$$\beta=\frac{\pi}{4l}=\frac{3.1416}{4(2330)}$$
$$=3.37\times 10^{-4}\text{ circular radian per centimeter.}$$

[3] W. L. Everitt, "Communication Engineering," second edition, McGraw-Hill Book Company, New York 18, N. Y., p. 169.

Using (55)

$$Z_o = \sqrt{Z_{oc}Z_{sc}} = \sqrt{629.5e^{-j59°.2} \times 629.5e^{j42°.8}}$$
$$= \sqrt{(629.5)^2 e^{-j16°.4}}$$
$$= 629.5e^{-j8°.2} = 623 - j89.8 \text{ (ohms)}.$$

Assume this length of line to be terminated in an impedance of

$$Z_R = 346.4 + j200 = 400e^{j30°}$$

Using (52)

$$Z_s = 629.5e^{-j8°.2} \frac{[346.4 + j200 + 623 - j89.8]e^{\alpha l + j\beta l} + [346.4 + j200 - 623 + j89.8]e^{-\alpha l - j\beta l}}{[346.4 + j200 + 623 - j89.8]e^{\alpha l + j\beta l} - [346.4 + j200 - 623 + j89.8]e^{-\alpha l - j\beta l}}$$

$$Z_s = 629.5e^{-j8°.2} \frac{[969.4 + j110.2]e^{\alpha l + j\beta l} - [276.6 - j289.8]e^{-\alpha l - j\beta l}}{[969.4 + j110.2]e^{\alpha l + j\beta l} + [276.6 - j289.8]e^{-\alpha l - j\beta l}}$$

$$Z_s = 629.5e^{-j8°.2} \frac{976e^{j6°.5}e^{\alpha l + j\beta l} - 401e^{-j46°.3}e^{-\alpha l - j\beta l}}{976e^{j6°.5}e^{\alpha l + j\beta l} + 401e^{-j46°.3}e^{-\alpha l - j\beta l}}$$

$$Z_s = 629.5e^{-j8°.2} \frac{976e^{\alpha l}e^{j(\beta l + 6°.5)} - 401e^{-\alpha l}e^{-j(\beta l + 46°.3)}}{976e^{\alpha l}e^{j(\beta l + 6°.5)} + 401e^{-\alpha l}e^{-j(\beta l + 46°.3)}}$$

$$Z_s = 629.5e^{-j8°.2} \frac{976e^{0.37}e^{j51°.5} - 401e^{-0.37}e^{-j91°.3}}{976e^{0.37}e^{j51°.5} + 401e^{-0.37}e^{-j91°.3}}.$$

From a table of values of e^x:

$$e^{0.37} = 1.4477 \quad \text{and} \quad e^{-0.37} = 0.6907$$

$$Z_s = 629.5e^{-j8°.2} \frac{1414e^{j51°.5} - 277e^{-j91°.3}}{1414e^{j51°.5} + 277e^{-j91°.3}}$$

$$Z_s = 629.5e^{-j8°.2} \frac{1414e^{j51°.5} + 277e^{j88°.7}}{1414e^{j51°.5} - 277e^{j88°.7}}.$$

Note above: $Ae^{j\phi} = -Ae^{j(\phi + 180°)}$

$$Z_s = 629.5e^{-j8°.2} \frac{880 + j1106 + 5.1 + j277}{880 + j1106 - 5.1 - j277}$$

$$Z_s = 629.5e^{-j8°.2} \frac{885.1 + j1383}{874.9 + j829}$$

$$Z_s = 629.5e^{-j8°.2} \frac{1640e^{j57°.4}}{1205e^{j43°.4}}$$

$$Z_s = 856e^{j5°.8} = 852 + j86.5 \text{ ohms.}$$

Discussion on

"Noise Figures of Radio Receivers"*

H. T. Friis

Dwight O. North:[1] Dr. Friis' article is a valuable contribution towards the standardization of techniques in receiver measurement and classification. His definitions of available power and of gain in available power will be especially useful, and should eventually become standardized terms in the lexicon. One wishes that his treatment of earlier definitions in this field had been accorded the same care.

Following reference to a paper of mine which defined and formulated certain concepts relative to the rating of noise in receivers,[2] he continues in reference to his own text. "In this paper a more rigorous definition of the standard of absolute sensitivity, the so-called noise figure, of a radio receiver is suggested." The assertion of greater rigor might happily be overlooked; yet his choice of labels is most confusing, for what he calls "noise figure" I had termed "noise factor," and I had employed "absolute sensitivity" in a quite different sense. In my work, and unlike noise factor, which is a numeric, absolute sensitivity referred to a field strength: that field strength, in a plane-polarized wave passing an antenna, necessary to produce a signal power at the detector of a receiver equal to the total noise power (from all sources) at the same point. I make this attempt to set the nomenclature straight because technical progress seems difficult enough even in the absence of verbal entanglement. As for "noise factor" versus "noise figure," there is no special plea from this quarter beyond the cogent observation that, while the term "noise factor" has found common usage in this country, in Britain its use appears to be exclusive.

The formulation of absolute sensitivity ran as follows, in practical units.

$$\overline{E^2} = \frac{240\pi^2}{\lambda^2} \cdot \frac{4kT_0\Delta f}{D^2(\Omega, \phi)} \left[F + \frac{T_a}{T_0} - 1 \right]$$

where E is field strength, λ is wavelength, k is Boltzmann's constant, T_0 is room temperature in degrees Kelvin, T_a is a fictitious temperature assigned to local space in recognition of the power of local noise fields, $D^2(\Omega, \phi)$ is the antenna's space-directivity function, F is the noise factor under review, and Δf is the over-all noise bandwidth.[3] The quantity in square brackets I termed the "operating noise factor," meaning the noise factor as modified through the use of a real rather than a dummy antenna, thus: $F_{op} = F + (T_a/T_0) - 1$. Neither the "operating noise factor" nor the "absolute sensitivity" has yet enjoyed general usage, partly through want of information as to the value to be assigned to T_a, partly because it is convenient and useful to compare receivers on the bench, for which purpose F itself was devised. But it should be thoroughly recognized that the noise factor alone is by no means a "standard of absolute

* Proc. I.R.E., vol. 32, pp. 419–422; July, 1944.

[1] RCA Laboratories, Princeton, New Jersey.

[2] D. O. North, "The absolute sensitivity of radio receivers," *RCA Rev.*, vol. 6, pp. 332–343; January, 1942.

[3] For a definitive description of T_a and $D^2(\Omega, \phi)$ see footnote reference 2.

sensitivity." That is, the improvement in performance represented by a reduction in noise factor of, say, several decibels *cannot be evaluated at all* without a knowledge of T_a. An appreciation of this fact is extremely important. So, also, are further studies to catalogue T_a at all wavelengths, at all seasons, and in all localities of technical interest.

A recent communication from the Radiophysics Laboratory, Sydney, Australia, emphasizes the need for revision of the definition of noise factor in a particular circumstance.[4] Noise bandwidth Δf was defined as the *over-all* bandwidth in my work; Dr. Friis' definition agrees. However, a superheterodyne receiver may possess a considerable response, between antenna and converter, at image frequencies. Other things being equal, and according to present definitions (if not usage), its noise factor might, in consequence of an augmented Δf, be quoted several decibels lower than the noise factor of a similar superheterodyne without image response! Before we decide to abandon this problem, let us agree to redefine Δf as the noise bandwidth in the *useful signal channel only*. In this event the receiver with image response yields a somewhat greater noise factor than one without image response—as it should, in view of the extra thermal noise of the antenna in the image channel; and, provided the local noise fields possess a uniform power distribution over the entire pass band, the revised expression for operating noise factor becomes

$$F_{op} = F + h(T_a/T_0 - 1)$$

where $h \equiv$ (entire noise bandwidth) $\div \Delta f$.
The formula for absolute sensitivity is revised only through F_{op}.

It is of great importance to understand this revision and the need for it, since a faithful adherence to the earlier rules for a determination of noise factor, when exercised in the measurement of a receiver with equal response in signal and image channels, would lead to a quotation some 3 decibels below its deserts.

Another limitation upon the concept of noise factor, this time constitutional, restricts its employ to that class of receivers which exhibits no measurably significant source of noise beyond the detector. For, the signal-to-noise ratio before a detector is transformed within the detector in a complex fashion which may depend upon the magnitude of both the signal and the noise. Fortunately the limitation is, on the whole, an academic one, since it is generally demonstrable that detection at low level is bad practice from a noise standpoint, and such designs can usually be avoided.

Of course, even with high-level detection, the signal-to-noise ratio ρ_2 *after* detection (by any reasonable definition) can be equal to the signal-to-noise ratio ρ_1 *before* detection only when the latter is large, and decreases more rapidly as ρ_1 approaches unity, finally assuming the form, $\rho_2 \propto \rho_1^2$, as $\rho_1 \rightarrow 0$. An understanding of this behavior is often a prerequisite to a comparison of alternative systems for a specific service,[5] and to studies of service quality or coverage. Even so, the subject would seem to fall outside the province of absolute sensitivity, whose scope is conceived to embrace ρ_1 alone. The relation between ρ_1 and service quality can be treated in no general terms.

For purposes of analysis, particularly in the event of nonuniform distribution of noise over the pass band, one often wishes to refer to a quantity associated with each differential element of bandwidth df, whose weighted average over the entire band is the noise factor. The weight factor is, of course, the gain in available power, or its equivalent, as employed in the definition of noise bandwidth. In accord with a proposal by E. J. Schremp,[6] we have found the term, "single-frequency noise factor," a definitive title for this quantity. Nevertheless, it lacks the brevity it merits; perhaps something better will eventually be suggested.

In order fully to convey a useful magnitude in the quotation of a noise factor, one must convey also the value of room temperature T_0 upon which it is based. The latter information is ordinarily suppressed. I indicated a preference[3] for 300 degrees Kelvin; Dr. Friis suggests 290. Of course neither of us is suggesting that the tester work in a thermostatted atmosphere of 81 degrees Fahrenheit or a comparatively chilly 63 degrees; nor are we even suggesting that the dummy antenna be maintained during test at a standard temperature. The requirement is merely that its temperature be ascertained and the computations be corrected to a standard temperature before a noise factor is quoted. It should be recognized that no such procedure need ordinarily be followed unless one's quotation carries a claim to accuracy of 0.1 decibel or better, except perhaps in unusual circumstances which subject the dummy antenna to a humanly uncomfortable climate.

H. T. Friis:[7] It is stimulating to read the discussion by Dr. North, who has contributed so much on receiver rating by his paper.[2]

The use of the phrase "more rigorous" in the first sentence of the second paragraph of the paper under discussion was perhaps unfortunate since I did not wish to imply that Dr. North's definitions were at all inaccurate, but merely that my suggested definitions were believed to be sufficiently precise to apply to four-terminal networks in general, including low-gain amplifiers and converters, and cases where a mismatch existed.

As far as the nomenclature is concerned, I am perfectly willing to leave the choice of terms up to the engineers who use the definitions and to those who will finally write the standards on the subject. The term

[4] Although unable to agree with their view that no need for revision exists, I am indebted to Drs. J. L. Pawsey and Ruby Payne-Scott of the Radiophysics Laboratory for pointing out this difficulty.

[5] For example, M. G. Crosby, "The service range of frequency modulation," *RCA Rev.*, vol. 4, pp. 349–371; January, 1940.

[6] Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.

[7] Bell Telephone Laboratories, Inc., Red Bank, New Jersey.

"noise figure," as is known, is widely used in this country and has appeared in at least one British publication.[8]

I used the words "absolute sensitivity" in their abstract sense. As defined by Dr. North or by me, the noise factor or noise figure of a receiver certainly is a measure of the absolute sensitivity of that receiver itself. I most certainly agree that the noise figure or factor is not a measure of the absolute sensitivity of a complete radio receiving installation. The importance of the local noise field in this case has long been recognized,[9] but so far we have not found it necessary or convenient to set aside any particular phrase to designate this absolute sensitivity. In radio-circuit design we use the noise-figure definition given by formula (5) in my paper for the four-terminal network made up of both the transmitting antenna and the receiving antenna, and we assign, as Dr. North has suggested, for the temperature of the receiving-antenna impedance, a value that takes care of the external noise sources. This way of rating a complete radio circuit is briefly outlined in my paper at the end of the section entitled "Measurement of the Noise Figure."

[8] D. K. C. MacDonald, "A note on two definitions of noise figure in radio receivers," *Phil. Mag.*, vol. 35, pp. 386–395; June, 1944.
[9] See papers by R. K. Potter, A. Reber, and K. G. Jansky in the PROCEEDINGS OF THE I.R.E. on atmospheric and other radio noise.

As Dr. North points out, it makes little difference whether the noise figure be defined for a temperature of 290 degrees Kelvin or 300 degrees Kelvin. I chose the value 290 degrees merely because it makes the value of KT a little easier to handle in computations.

I am glad to see that Dr. North calls attention to the effect of image response on the noise figure. For reasons of clarity I made my paper as brief as possible and a great many details have therefore still to be settled. He has also pointed out that the concept of noise figure cannot be applied indiscriminately to the final or audio detector of a receiver. Fortunately, as he has stated, the limitations are such that they will have little effect upon the usefulness of the noise figure as a standard.

Television Awards

Sixteen leaders in the television field were presented awards on December 11, 1944, for outstanding contributions to television development at the First Annual Conference Banquet of the Television Broadcasters' Association at the Commodore Hotel ballroom in New York City.

The presentations were made before a gathering of more than 1000 by Paul Raibourn, president of Television Productions, Inc., and in charge of television activities for Paramount Pictures, Inc., who was chairman of the Committee on Awards, which included Frederick R. Lack (A'20-F'37), and Orestes H. Caldwell (M'40-SM'43). Each recipient was presented a gold medal.

These awards were the first ever bestowed for achievement in the advancement of the art, science, and industry of television. They were made in three classes, including techical achievement, program achievement, and general achievement in television. Those made to I.R.E. members are listed below:

First Award

Dr. Vladimir K. Zworykin (M'30-F'38), RCA Laboratories, Princeton, N. J., with the citation: "For development of the iconoscope and the storage principal of picture pickup, resulting in the first practical television pickup equipment."

Awards for technical pioneering in television were made to:

Philo T. Farnsworth (A'28-M'34-F'39), Farnsworth Radio and Television Corporation, Fort Wayne, Ind., with the citation: "For work on television scanning methods and the electron multiplier." A similar award was made to Lloyd Espenschied (M'13-F'24), Bell Telephone Laboratories, New York City, with the citation: "For adapting the coaxial cable to transmitting wide bands of radio frequency suitable for modern television."

Dr. Peter C. Goldmark (A'36-M'38-F'42), Columbia Broadcasting System, New York City, received an award with the citation: "For work in the development of motion-picture pickup equipment and electronic analysis and control of equipment for color television." And F. J. Bingley (A'34-M'36-SM'43), Philco Radio and Television Corporation, Philadelphia, Pa., received an award with the citation: "For improvement in contrast of television pictures through flat-face tubes and experiments on link operations particularly as regards outdoor events."

A technical pioneering award was also granted to Dr. Allen B. DuMont (M'30-F'31), Allen B. DuMont Laboratories, Passaic, N. J., with the citation: "For the development of the cathode-ray tube to a satisfactory commercial instrument of television control and reproduction."

An award was made for general contribution to television to each of the following and this award also does not cover the past year, but it is a summation of the efforts of many years.

First Award to General David Sarnoff (A'12-M'14-F'17), former Secretary of The Institute of Radio Engineers and on leave from the Presidency of Radio Corporation of America, with the citation: "For his initial vision of television as a social force and the steadfastness of his leadership in the face of natural and human obstacles in bringing television to its present state of perfection."

Other awards of this class were made to Dr. W. R. G. Baker (A'19-F'28), with the citation: "For his leadership in standardizing television through the National Television System Committee and supporting it through the Radio Technical Planning Board"; to David B. Smith (A'35-SM'44), Philco Radio and Television Corporation, Philadelphia, Pa., with the citation: "For his work on the National Television System Committee and his planning of television future as panel chairman with the Radio Technical Planning Board"; and to Dr. Alfred N. Goldsmith (M'12-F'15), Editor of the PROCEEDINGS of the I.R.E., with the citation: "For his work on the National Television System Committee and the Radio Technical Planning Board and his vision of the relationship of the motion picture and television."

Books

Building-Fund Campaign Launched

BY THE time this article appears in the PROCEEDINGS, the Building-Fund campaign to raise not less than $500,000 will have been formally launched at the Winter Technical Meeting, and all the Sections will have been informed as to their participation. The objective is to secure, during February and March, donations to the Building Fund from every member and friend of the Institute, including friendly corporations. To attain this objective, the campaign organization has been so designed as to insure that personal calls will be made by members, usually in pairs or groups, upon all members and upon certain corporations in the United States and Canada, to set forth fully our need and our plans, and to facilitate the making of contributions. Reliance will be placed upon telephone or direct-mail appeals only in the case of foreign members and others too remotely located to make personal calls feasible.

WHY $500,000 OR MORE?

Months of study, conference, and investigation have culminated in this action. Members will recall from previous PROCEEDINGS articles, that the Board of Directors initiated this campaign for the financing of "a suitable headquarters building, whether alone or in association with other engineering societies as the opportunity presents." With this aim in mind, the entire Building-Fund program has been and will be kept flexible. Essential to final determination of detailed plans, however, is the primary objective: successful financing at the earliest possible date.

The fact that the goal has been set at the sum of not less than $500,000 does not imply that we seek to acquire a half-million-dollar building. If the cost of land and building, including alterations and furnishings, were placed at $300,000, wisdom would dictate the inclusion of an additional $200,000 representing the capitalized difference between the present office rent and the estimated annual maintenance of the building. The Board feels that maintenance to the extent of present rent is an obligation to be met out of current income, and that excess maintenance obligations taken on with a building should be capitalized and the amount raised with the fund collected to buy or build. The Board also feels that it has no right to ask its members and corporate and other friends to contribute to a sinking fund to cover depreciation, so no amount for depreciation is included. By this campaign, this generation plans only for itself —a matter of, say, 25 to 40 years—knowing that our successors, alert to changed conditions, will in their time probably initiate more ambitious plans.

The amount of $500,000 is also felt to be adequate to provide freedom of action alternatively to join with other engineering and scientific societies in occupying acquired premises, if to do so proves ultimately more feasible and desirable. If, for example, favorable consideration were to be obtained from one of several foundations, or by other means, to establish an Engineering and Science Center, realization of the amount which is our goal would put the Institute in the same financial company as other societies likely to be included in any grant.

A PROGRAM OF EXPANSION

If the radio-and-electronic field held less brilliant promise for the future; if the growth and prestige of I.R.E. were unlikely to keep step with the expansion of its field; if its Directors had no intention and no plans for giving increased service to its membership and to the industry—there would be no justification for a fund to house the Institute as other similar societies are housed. It is of utmost importance to remember that *the prospective housing of our activities is merely a necessary means to their expansion*. Therein lies the real interest of every contributing member and corporation.

Through wise stewardship, successive Boards of Directors have built up the Institute's reserves and surplus so that we do not have to go before our prospective corporate contributors as beggars, but as a body of engineers with a respectable financial standing and record of good management, seeking funds to broaden our activity and usefulness. If the Directors had felt it desirable, they could have allocated the first $100,000 to the Building Fund from invested funds of the Institute. But they realize that a real program of expansion will take money for many purposes besides that of a building, its furnishings, and its maintenance.

The present funds of the Institute have accordingly been earmarked, not for the Building Fund, which is to stand on its own feet, but for the greater service of the Institute to its membership and the industry. Corporate contributors to the Building Fund, who have a right to expect that the stewardship basis of a request for a corporate contribution shall be as valid as that for a loan from a bank, will be impressed by the judgment of the Directors in not obligating, for a building, the funds now in its investment account.

Although this plea for donations comes before the membership as a Building Fund, from the foregoing it will be recognized that raising of the Fund is only part of a program far broader than that: rather, it is the first of a series of steps (and being the most expensive, the only one we cannot financially handle without the assistance of our friends) to put The Institute of Radio Engineers where it belongs—in the forefront of engineering, alongside other societies which give their respective memberships and industries better service than I.R.E. up to now has had office space and staff enough to

at Winter Technical Meeting

render. Because it is the first step towards enlarged service, the Building Fund is the Institute's great opportunity. To succeed in its realization presents a challenge much greater than any that has previously been before us. Upon the way that we accept it will undoubtedly depend much of the entire future course of our society.

Success Depends on Membership

The outcome of the Building-Fund campaign necessarily rests with the membership of the Institute. The Directors have engaged competent fund-raising counsel in order to promote efficiency and to save time and money; and, because there is no room at Institute headquarters, have temporarily rented campaign offices in Suite 930 at 55 West 42nd Street, New York. The Board and its committees are doing their utmost to promote efficient planning and comprehensive presentation of this important project. All this will be futile, however, without the full co-operation of the members in financial support, and by assistance in the securing of subscriptions.

Appeals will be made, for the most part, through Section Building-Fund Committees organized in the various Sections, including the New York Section (the Canadian Sections are to organize under the Canadian Council). The chairman of each of these Section Building-Fund Committees: (a) will be appointed by the Chairman of each Section; (b) will receive instruction from and be responsible to the Sections Solicitation Committee chairman at campaign headquarters; (c) will himself appoint and direct two line committees: (1) a Membership Solicitation Committee to direct the activities of pairs or groups calling on members; and (2) a Corporation Solicitation Committee to direct calls made on officers of selected corporations; plus five staff committees (wherever the Section is large enough to justify), to handle: (3) prospect ratings and quotas, (4) accounting and audit, (5) local publicity, (6) training and meetings, and (7) liaison with members not on the Section list but in the geographical vicinity, and with college representatives in the vicinity.

An Initial Gifts Committee will function at campaign headquarters, its chairman making direct appointment of individuals and teams to make the approach to the larger corporations and individual givers wherever located in the United States and Canada. These firms and individuals will be "blocked" against duplicate solicitation by Section Building-Fund Committees, but donations secured from them will be credited to the proper Sections.

Section Building-Fund Committees will be furnished all necessary further information on organization, procedure, handling and accounting for funds, and so on. Building-Fund representatives will, so far as possible, visit the various Sections during February and March.

A One-Time Appeal

To guarantee that funds paid in by subscribers shall be properly safeguarded, and that they be disbursed for the designated purposes only, a suitable agreement has been executed, particulars of which will be made available to all potential donors. Assurances that gifts to the Building Fund are deductible for Federal Income Tax, Estate, and Gift Tax purposes, have been secured from the Commissioner of Internal Revenue, particulars of which will also be made available.

Because a Building-Fund campaign is nonrecurring, has not been made previously in the engineering profession for approximately 40 years, is not likely to be made by the Institute again during the lifetime of most of the present members, gifts may be "stepped up" higher than customary in the case of annually recurring "drives" for charity. The fact that the net cost to the donor for the gift is a fractional amount of the face of the checks drawn (by the percentage of taxes that would have to be paid were no gift made), is a further justification of liberality. Notes on pledges may be given to spread the amounts over 1945 and early 1946 to suit the donor's financial and tax program. If full advantage is taken of these three inducements to liberal giving, there will be no doubt of a happy conclusion to the Building-Fund campaign.

The Case for the Building Fund

In summary of this article and the two which preceded it in the December and January issues of the Proceedings, the following statements appear to be axiomatic:

(a) *Postwar need for I.R.E. will be greater than ever before.*

(b) *The Institute must have more room to function efficiently.*

(c) *Vision is needed if the housing problem is to be solved adequately.*

(d) *Support must come from the Institute's friends.*

(e) *Now is the time for action.* The need for sufficient space is not theoretical; the overcrowded offices are a reality today. Delay will only permit the problem to be aggravated as tomorrow's need for expansion crowds in upon us. More temporary movings will disrupt operation and leave us no better off than we are now. This situation calls for a permanent solution; and the Board believes all members will agree that an aggressive Building-Fund campaign is the answer.

Institute News and Radio Notes

Board of Directors

November 29 Meeting: At the regular meeting of the Board of Directors, which was held on November 29, 1944, the following were present: H. M. Turner, president; R. A. Hackbush, vice-president; S. L. Bailey, W. L. Barrow, E. F. Carter, I. S. Coggeshall, W. L. Everitt, president-elect; Alfred N. Goldsmith, editor; R. F. Guy, R. A. Heising, treasurer; Keith Henney (guest), F. B. Llewellyn, Haraden Pratt, secretary; H. J. Reich, B. E. Shackelford (guest), H. A. Wheeler, L. P. Wheeler, W. C. White, and W. B. Cowilich, assistant secretary.

Constitutional Amendments

Report: The November 27, 1944, report of the Tellers Committee, was unanimously accepted and the amendments approved by the voting membership were declared adopted.

Mr. Heising, chairman of the Constitution and Laws Committee, pointed out that the following proposed amendments failed to pass:

No. 2—The purpose of this amendment was to substitute the names of Member, Associate, and Affiliate for the present names of Senior Member, Member, and Associate, respectively.

No. 6—The purpose of this amendment was to increase the dues.

No. 10—This amendment was a limitation upon the appointment of Secretary, Treasurer, Editor, and appointed members of the Board.

Second Constitutional-Amendment Ballot: This ballot, to include the amendment of Article IV recently submitted by petition and relating to a second plan to increase membership dues, should be mailed after the annual meeting of the Sections Committee scheduled for January 24, 1945. It was noted that H. P. Westman, who initiated the petition, is agreeable to the postponement.

In the publicity on the foregoing dues-increase plan, it was considered important to emphasize that the resulting larger revenue would make possible more service to the membership and an increase in Section rebates. The suggestion was also made to the effect that the Sections should be given ample opportunity to discuss this plan and requested to report on such discussions to the Institute.

Following the discussion, the following motion was unanimously approved:

"Because the proposed amendment will primarily affect the new members in 1945, the Constitution and Laws Committee is empowered to decide when the petitioned amendment shall be submitted to the voting membership before July 1, 1945."

Bylaws: These proposed bylaws sections, which had been considered at recent meetings, were approved:

Section H. "The Board, by this Bylaw, waives the dues of each and every member of the Institute who has attained the age of 65 years and has been a member of the Institute for 35 years."

Section B. "The PROCEEDINGS shall be sent to all members. All members, except Students, shall be entitled to a YEARBOOK, if it is published. The Board of Directors shall determine the distribution of other publications."

Awards

Medal of Honor for 1945: This citation for H. H. Beverage, to whom the medal was awarded at the last meeting, was given unanimous approval:

"In recognition of his achievements in radio research and invention, of his practical applications of engineering developments that greatly extended and increased the efficiency of domestic and world-wide radio communications, and of his devotion to the affairs of The Institute of Radio Engineers."

Morris Liebmann Memorial Prize for 1944: W. W. Hansen was chosen by the committee as the recipient for the 1944 prize, with the following citation:

"For application of electromagnetic theory to radiation, antennas, resonators, and electron bunching, and for the development of practical equipment and measurement techniques in the microwave field."

Fellowships: 12 candidates were elected to the grade of Fellow and approval was given to their citations.

H. H. BUTTNER—"In recognition of his direction of radio communication activities in the international field."

O. H. CALDWELL—"For his contribution in broadening the horizon of the engineer by his long-continued effort to increase the use of electronic principles in industrial operations."

W. H. DOHERTY—"For his contributions to the development of radio transmitting equipment."

A. W. HULL—"In recognition of his many contributions to the development and design of electron tubes both for radio and industrial applications."

A. L. LOOMIS—"In recognition of his work in the application of electronic techniques to medical research and for contributions to microwave development."

A. V. LOUGHREN—"For his many valuable contributions to broadcast and television engineering and his untiring efforts to advance this profession."

F. X. RETTENMEYER—"In recognition of his accomplishments in the development of broadcast and automobile receivers and aviation radio."

S. A. SCHELKUNOFF—"In recognition of his mathematical contributions to electromagnetic theory."

R. L. SMITH-ROSE—"In recognition of his pioneer work in the field of direction finding and radio propagation, allied to his leadership of an outstanding radio research group."

K. S. VAN DYKE—"In recognition of his work in research on characteristics of piezoelectric crystals and their application to frequency control."

E. M. WEBSTER—"For his contributions to the development of the maritime mobile radio services and his leadership in promoting measures for enhancing the safety of life and property at sea."

P. D. ZOTTU—"For his contributions in the field of high-frequency heating, particularly in the application of dielectric heating in industry."

Sections

San Diego: The establishment of an Institute Section at San Diego, California, recommended by the Executive Committee, was authorized with San Diego County as the official territory.

Cedar Rapids: Mr. H. A. Wheeler reported that the November 25, 1944, petition for the formation of an Institute Section at Cedar Rapids, Iowa, contained the signatures of thirty qualified members, or five more than the number prescribed in the Bylaws, and thus is in good order.

After considering the petition, it was moved to authorize the establishment of the Cedar Rapids Section with the official territory to consist of the thirty-two counties which are within a sixty-mile radius of the Section.

COUNTIES IN IOWA

Bremer	Marshall	Henry
Fayette	Benton	Johnson
Clayton	Keokuk	Cedar
Grundy	Jefferson	Clinton
Black Hawk	Linn	Scott
Iowa	Jones	Mahaska
Louisa	Jackson	Muscatine
Delaware	Jasper	Tama
Buchanan	Poweshiek	Polk
Dubuque	Washington	Story

COUNTIES IN ILLINOIS

Rock Island	Mercer

Committees

Code of Ethics: Upon recommendation of the Executive Committee, a special Committee on Code of Ethics was created and W. L. Barrow, chairman; Alfred N. Goldsmith, and L. E. Whittemore were named members of the group.

Technical: The following persons, including those recommended by the Executive Committee, were appointed.

ELECTRONICS
G. W. Greer

SYMBOLS
H. F. Dart

TELEVISION
R. N. Harman-H. T. Lyman

Standardization of Program Loops and Private-Line Terminations: Dr. Llewellyn, in charge of standardization and other technical committees, referred to the recent letter from Edward J. Content, suggesting the formation of a committee on the indicated subject to include two representatives of the Institute. This matter was considered at the recent Executive Committee meeting.

H. A. Chinn and O. B. Hanson were named to serve as the Institute's representatives on the committee, which is to study the standardization of program loops and private-line terminations at points of origin of radio programs.

Rochester Fall Meeting: The November 27, 1944, letter from Mr. Horle, containing a report on the named meeting, was read by President Turner and it was noted that the registered attendance was more than 700.

President Turner and President-Elect Everitt, who attended the meeting, stated that the meeting was regarded as successful.

1945 Summer Convention: Mr. H. A. Wheeler, the Executive Committee member responsible for conventions and conferences, gave reports on matters concerning this convention.

It was unanimously decided to hold this convention at Montreal, Canada, approximately June, 1945, and to communicate with the Montreal Section relative to making preliminary arrangements for the convention.

The Institution of Electrical Engineers: Upon recommendation of the Executive Committee, these members were appointed to the Liaison Committee: Ralph Bown, *chairman;* F. S. Barton, and F. B. Llewellyn.

Office Organization: Secretary Pratt explained the proposed organization chart and called attention to the necessity for taking immediate steps to fill the three new positions listed below:

Executive Secretary
Technical Editor
Technical Secretary

The following motions were unanimously approved:

"The Board establishes the positions of Executive Secrdtary, Technical Editor, and Technical Secretary, and authorizes that provisions be made for these positions in the 1945 budget."

"It was moved to appoint a committee consisting of S. L. Bailey, *chairman;* F. E. Terman, W. O. Swinyard, F. R. Lack (Ralph Bown, *alternate*), the Secretary, and the Editor, for the purpose of entering into negotiations without making commitments with prospective candidates, and to instruct this committee to submit their recommendations to the Board on candidates for the positions of Executive Secretary, Technical Editor, and Technical Secretary."

It was stated that these additions to the staff are intended to broaden future Institute activities, including the expansion of the publications program and the increase in membership. It was also indicated that a deficit could be expected for a few years from the Institute's organization and activities on the new basis.

President's Traveling Expenses: Upon the recommendation of the Executive Committee, the sum of $1000 was appropriated for the traveling and related expenses of the President of the Institute, the appropriation to become effective in 1944 and to continue annually thereafter until further notice.

H. J. van der Bijl: President Turner called attention to this radiogram, addressed to Secretary Pratt and received on November 28, 1944, from Vice-President-Elect van der Bijl relative to his recent election to the stated office:

"Please convey to Institute my sincere appreciation electing me vice-president 1945 which I cherish as exceptional honor. My best wishes for continued success and evergrowing importance of contribution to civilization by its Fellows and members."

Societé des Radioelectriciens: This message, dated November 27, 1944, from Monsieur Bouthillon was read by by Secretary Pratt:

"The Societé des Radioelectriciens at its meeting of November 25th held at the Sorbonne, first meeting since March, 1940, voted that a message be sent to The Institute of Radio Engineers to mark and celebrate the occasion of renewed co-operation with the American radio scientists and engineers through the Institute. It is the firm belief of the Societé des Radioelectriciens that proper co-operation with the U. S. Radio Institute will help materially in the common cause. Signed, Bouthillon, President of Societé des Radioelectriciens."

The Board's appreciation of the message was expressed and Secretary Pratt was requested to acknowledge the communication.

It was also decided to bring a copy of the message to the attention of the Liaison Committee, which is in the process of formation.

Executive Committee

November 27, 1944: The following members were present at the November 27, 1944, meeting of the Executive Committee: H. M. Turner, president; W. L. Everitt, president-elect; E. F. Carter, Alfred N. Goldsmith, editor; R. A. Heising, treasurer; F. B. Llewellyn, Haraden Pratt, secretary; H. A. Wheeler, and W. B. Cowilich, assistant secretary.

Membership: The following applications for membership were approved: for transfer to Senior Member grade, C. J. Bridgland, J. H. Clark, J. D. Cobine, R. W. Deardorff, Marcy Eager, Leonard Everett, Harold Goldberg, W. S. Marks, Jr., Eugene Mittelmann, H. T. Pekin, L. C. Shapiro, A. G. Simson, and R. F. Walz; for admission to Senior Member grade, R. W. Larson and W. P. West; for transfer to Member grade, J. J. Adams, W. G. Baptist, R. B. Bonney, A. F. Brewer, A. W. Burks, P. J. Konkle, F. T. Mitchell, Jr., B. E. Montgomery, E. W. Novy, J. J. Okrent, William Ostaff, V. L. Palmer, C. J. Probeck, O. W. B. Reed, Jr., and K. E. Reese; for admission to Member grade, A. R. G. Albright, W. M. Bauer, P. L. Bennett, B. L. Griffing, R. A. McNaughton, Elliott Mehrbach, A. H. Meyerson, L. F. Millett, F. R. Park, G. N. Patchett, L. M. Rundlett, R. E. Stemm, C. R. Stoner, and F. P. Thomson; Associate grade, 185; and Student grade, 126.

Constitutional-Amendment Balloting

Of the 3660 ballots on Constitutional Amendments, dated September 15, 1944, and mailed on that date to the voting membership, 1742 were received in good order, according to the report of the Tellers Committee.

The Constitution prescribes that in such balloting, at least 20 per cent of all voting members must participate and a minimum of 75 per cent of the votes cast must be affirmative to adopt a proposed amendment.

The seven of the ten proposed amendments, approved by the voting membership, were adopted by the Board of Directors at its meeting on November 29, 1944 and by that action became effective on December 29, 1944.

The table at the foot of this page shows the results of the balloting.

The amendments, including those approved and not approved, are listed below and in each case the wording, or the explanation of the purpose, is indicated:

No. 1 (Approved)

Article I, Section 2

Insert after the first word "be" the following words "scientific, literary, and educational. Its aims shall include" making it read

"Section 2. Its objects shall be scientific, literary, and educational. Its aims shall include the advancement of the theory and practice of radio, etc."

No. 2 (Not Approved)

The purpose of this amendment was to substitute the names Member, Associate, and Affiliate for the present names of Senior Member, Member, and Associate, respectively.

No. 3 (Approved)

Article II, Section 1d

Delete the last sentence, "Furthermore, etc."

Proposed Amendments	Votes in Good Order	Per Cent of Votes Cast	Votes Required to Approve	For	Against
No. 1	1665	46	1249	1562	103
No. 2	1712	47	1284	1022	690
No. 3	1686	46	1265	1532	154
No. 4	1685	46	1264	1606	79
No. 5	1703	47	1278	1603	100
No. 6	1716	47	1287	1055	661
No. 7	1713	47	1285	1666	47
No. 8	1715	47	1287	1538	177
No. 9	1716	47	1287	1676	40
No. 10	1695	46	1272	649	1046

No. 4 (Approved)

Article II, Sections 2 and 3

Delete the word "either" in first sentence of both sections.

Article II, Section 2, change periods to commas and add word "or," at end of paragraphs *a*, *b*, and *c*.

Article II, Section 3, change periods to commas and add word "or," at end of paragraphs *a*, *b*, and *c*.

Article II, Section 4, change period to comma and add word "or," at end of paragraph *a*.

No. 5 (Approved)

Delete in Article II, Section 1*e*, everything after the word "Student" and substitute

"who may participate in meetings, wear the badge of the Institute, and receive publications designated by the Board of Directors, but who shall have no other rights and privileges."

No. 6 (Not Approved)

The purpose of this amendment was to increase the dues.

No. 7 (Approved)

Article VII, Section 1

Insert in the second paragraph, second sentence, between the words "before" and "August" the words "twelve o'clock noon on the last week-day prior to" making it read "For acceptance a letter of petition must reach the executive office before twelve o'clock noon on the last week-day prior to August fifteenth of any year, etc."

In the fifth paragraph, fourth sentence, insert between the words "office" and "prior" the words "before twelve o'clock noon on the last week-day" making it read "Only ballots arriving at the executive office before twelve o'clock noon on the last week-day prior to October twenty-fifth shall be counted."

No. 8 (Approved)

Article IX

Delete title and both Sections 1 and 2. SUBSTITUTE—SECTIONS AND OTHER GROUPS. Section 1. The Board of Directors may authorize the establishment of sections and other groups of members for the purpose of promoting the interests of the Institute. The Board of Directors may, at its discretion, terminate the existence of any such group.

No. 9 (Approved)

Article X, Section 2

Insert after the first sentence—

"The ballots after marking shall be placed in plain sealed envelopes, enclosed within mailing envelopes marked 'ballot' and bearing the member's written signature. Only ballots within signed outer envelopes shall be counted. No votes by proxy shall be counted. Only ballots arriving at the executive office prior to the stated time limit shall be counted."

No. 10 (Not Approved)

Article VII, add:

"Section 4. No person shall be eligible for appointment by the Board as Director, Secretary, Treasurer, or Editor after having accepted five such appointments to any or several of these offices."

London, Ontario, Section Formed

An instrument, still on the secret list, actually launched the London, Ontario, section of The Institute of Radio Engineers at the University of Western Ontario on November 24, 1944, by carrying the image of the charter of the section from one room to another. The layman might say the section and its charter were projected into being on the screen of a little black case—contents unknown.

Culmination of a year's organization work marked the charter night of the London section of the parent organization designed to advance and co-ordinate discoveries and methods of furthering electronics and electrical communication through discussion of individual and group problems.

Wing Commander K. R. Patrick, officer commanding the R.C.A.F. station at Clinton, and several of the camp personnel handled the equipment which projected the London Section's charter on a screen from another room by means of a new departure in television. Flight Lieutenant Robert Wilton, secretary of the new section, then read the charter from the screen.

Election of a permanent slate of officers, reading of congratulatory messages, and a technical description of Canadian and American radio planning, marked the charter ceremonies. Ralph W. Hackbusch, vice-president of The Institute of Radio Engineers, explained the organization of Radio Technical Planning Boards both in the United States and in Canada, and stated that advancement of radio and electronics through these groups is assured.

Dr. Sherwood Fox, president of the University of Western Ontario, welcomed the members of the London section and congratulated them on receiving their charter. Dr. Fox offered any available facilities of the university to the London section for future meetings, and use of any equipment that might be of assistance in furthering the work of the members.

Congratulations and greetings were offered by E. B. Buchanan, general manager of the London Public Utilities, and Charles Miller, of the Ottawa section of the I.R.E. And so, in striking and appropriate fashion was started the promising career of the new section.

AFTER CEDAR RAPIDS MEETING

R. V. Guettler, G. Milton Ehlers, and W. S. Parsons, all of Centralab Company; Seated, T. A. Hunter, Collins Radio Company, Temporary Chairman, Cedar Rapids Section.

Cedar Rapids Section Formed

On November 29, 1944, 64 people met in Cedar Rapids, Iowa, to organize a proposed new Section of The Institute of Radio Engineers. The section is to include thirty counties in Iowa and two counties in Illinois, covering approximately a sixty-mile radius from Cedar Rapids.

A paper on "The Romance of Ceramics" was presented by G. Milton Ehlers, head of the ceramics division of the Centralab Company, Milwaukee, Wisconsin; and R. V. Guettler, of the silver mica division at Centralab, also spoke briefly on the subject of silver mica condensers.

It is estimated that the new section will have a charter membership of approximately 100; the election of permanent officers was scheduled for the next meeting. Temporary officers appointed were T. A. Hunter, Chairman, and John A. Green, Secretary-Treasurer.

Following the meeting, those in attendance were invited by W. S. Parsons, assistant sales manager of the Centralab Company, to be guests of that organization at a luncheon.

Since this meeting, the Board of Directors of the Institute has formally authorized the Cedar Rapids Section, for which a successful and constructive career is hoped and anticipated.

Research Scientist Views Electronic Future

Dr. O. S. Duffendack (SM'44) research director of North American Philips Company, Inc., stated in a recent talk before the Yonkers, New York, Chamber of Commerce that although progress along some lines in electronics has been phenomenal and that a very noticeable effect on our mode of life will certainly be evident in the decade following this world war, it was necessary to sound a note of caution.

He said, "Don't expect miracles. What has happened is that certain developments—not all—in this field have been speeded up so that we shall have available at the end of the war, devices and processes that normally would not have been ready for another ten or possibly twenty years. To anyone who knows research and development, that is almost a miracle.

"Without violating the rules of military security we may predict some of the changes that may be anticipated as a consequence of electronic developments. Radio communication especially will benefit from the availability of higher frequency bands, and marked improvement in frequency-modulation techniques. New tubes, new conductors, new receivers, and new circuits will make possible a greater use of radio communication without mutual interference. We may expect to see a greater use of long-distance and localized services as well as in local police-radio systems. Fleets of trucks may be directed by radio and plane-to-plane, train-to-train, as well as engine-to-caboose communication by radio, wire, or rail are probable. The improvements in small portable two-way sets, such as the walkie-talkie, have been rapid; and a widespread use of these devices by surveying, prospecting, and exploring parties is certain.

"Improvements in antenna design and other things have resulted in sharper, better defined radio beams that will be reflected in a marked increase in beam communication and in improved air navigation. The development of quartz-crystal oscillators has been truly phenomenal. Now a single plant produces in a single month *more* crystal oscillators of *higher* quality than were produced *in the entire world in a year before the war*. As a consequence, close frequency control and precise tuning are making possible narrower channels and improved radio operation. The story of the development of processes for the large-scale manufacture of crystal oscillators is the fascinating one of an achievement for which the electronic industry can be very proud.

"Broadcasting and radio communications likewise will be noticeably improved and extended by developments in the high-frequency ranges and frequency modulation. Television is certain to have a rapid spread to continental coverage in a few years after the war. It was ready for this when the war broke. Color television is a definite possibility and will follow in due course. Facsimile transmission, which is not a specific war development, likewise will benefit from others and be quickly improved.

"Transportation will be faster and safer as a consequence of wartime electronic developments. Transoceanic flights were barely emerging from the stunt and novelty stage when the war began. Now that thousands of planes have been flown across both the Atlantic and the Pacific, transoceanic traffic by air is commonplace with a remarkably low rate of loss. We can give due credit to automotive and aerodynamic developments and still say that without new and improved navigation aids resulting from electronic research, the present achievements would have been impossible.

"Radar is the outstanding achievement in electronics during the war. After the war, not a passenger ship will sail nor a transport plane take off without the benefit of navigation aids from some type of radar making passage safe which before would be hazardous. Radar and radio, underwater sound, and other devices will guide and protect our ships of the sea and of the air and bring them safe to port through darkness and storm and hail and fog. Blind landing of airplanes will be commonplace and the mail will go with passengers through storm and cloud.

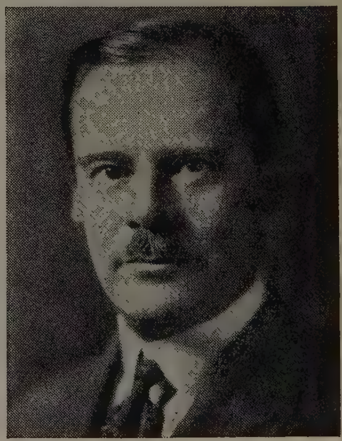
"Buzz bombs and rockets and V-2's are instruments of terror and death in this war. Now they carry explosives and we dread their approach. But in times of peace, these same carriers can be tamed and used for our benefit. Electronic devices to guide them and control their landing are entirely within the realm of the possible so that instead of sending death and destruction to London we may send gifts and benefits. The world again will shrink and become small through the achievements of electronics and they will find that the lives of our children will be filled with broader and fuller experiences than our own."

Dr. Duffendack, formerly professor of physics at the University of Michigan, is best known for his work in the field of electric conduction through gases and has published a number of scientific papers.

I.R.E. People

E. F. W. Alexanderson

The Edison Medal for 1944 has been awarded by the American Institute of Electrical Engineers to Doctor Ernst Fredrik Werner Alexanderson, consulting engineer, General Electric Company, "For his outstanding inventions and developments in the radio, transportation, marine, and power fields." It is awarded annually for "meritorious achievement in electrical science, electrical engineering, or the electrical arts" by a committee composed of twenty-four members of the A.I.E.E.



E. F. W. Alexanderson

Dr. Alexanderson joined the Institute of Radio Engineers as an Associate in 1913, transferred to Member grade in 1913, and to Fellow grade in 1915.

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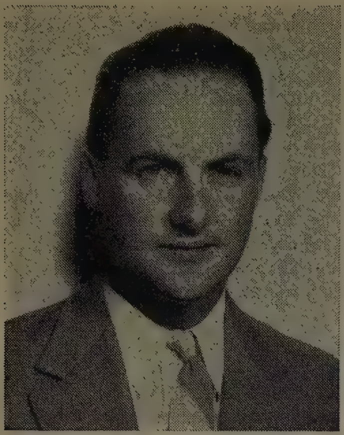
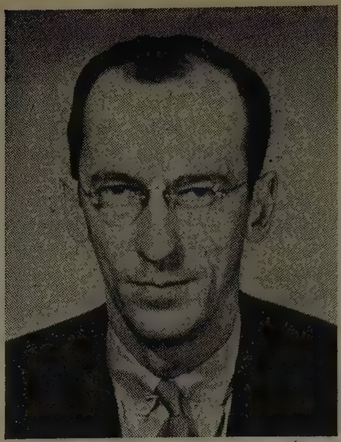
John M. Miller, Jr.

John M. Miller, Jr. (A'41), has been made chief engineer at United Cinephone Corporation, Torrington, Conn., in which capacity he is in charge of the corporation's designing and development work. Mr. Miller formerly did radio design and development work with Philco, the Navy Department, and RCA Victor, and is a member of the Circuits Committee of the I.R.E.



John M. Miller



William W. Garstang

Roger M. Wise

William W. Garstang

William W. Garstang (A'31-M'35-SM'43), is vice-president and director of Electronic Laboratories, Inc., of Indianapolis, Ind. Under his direction there are produced light- and heavy-duty vibrators and vibrator power supplies applicable to a wide range of devices from smallest portable radio equipment to fluorescent lighting in vehicles, and also permitting the operation of refrigerators and other appliances for nonstandard power supplies.

Winfield G. Wagener

Winfield G. Wagener (A'29-SM'44), formerly chief engineer at Heintz and Kaufman, has been appointed chief engineer of the vacuum-tube division of Litton Engineering Laboratories, Redwood City, Calif.

Graduated from University of California in 1928 with honors, Mr. Wagener won the John W. Mackay scholarship for advanced work in electrical engineering and, with its aid, continued to his master's degree. He also engaged in research work at Stanford under former I.R.E. President, Dr. F. E. Terman. At the present time he is chairman of the San Francisco section of I.R.E. Mr. Wagener is author of "Calculating the Performance of Vacuum Tubes," included in many standard books on radio. He has originated a number of practical applications of vacuum tubes in ultra-high-frequency circuits.

Prior to his previous connection he spent several years in development work with Federal Telegraph Company and later served with the Radio Corporation of America.

Bennett S. Ellefson

Bennett S. Ellefson (A'38-M'44) has been appointed assistant to the vice-president in charge of engineering of Sylvania Electric Products, Inc. Associated with the company since 1937, Dr. Ellefson has specialized in research on fluorescent screens, special uses of glass, fluorescent powders for cathode-ray tubes, and specialized war products. He holds a number of United States and foreign patents on fluorescent materials and glass, including certain special war products.

Prior to joining Sylvania, Dr. Ellefson conducted important research on industrial chemical processes. He is a member of the American Society for the Advancement of Science, the American Chemical Society, the American Ceramic Society, Phi Kappa Phi and Sigma Xi. A native of Canby, Minnesota, he holds a B.A. degree from St. Olaf College, an M.S. degree from the University of Minnesota, and a Ph.D. degree from Pennsylvania State College.

Roger M. Wise

The appointment of Roger M. Wise (A'26-M'30-F'37) to the newly created post of vice-president in charge of engineering of Sylvania Electric Products, Inc., has been announced by Walter E. Poor, the company's president.

Sylvania's director of engineering for the past two years, Mr. Wise previously served as the company's chief radio engineer for ten years.

Prior to joining the company in 1929, Mr. Wise was associated with the Remler, Cunningham, and Grigsby-Grunow companies.

After serving in the United States Navy during World War I as chief electrician (radio), Mr. Wise completed his education at the University of California.

Winfield G. Wagener

Bennett S. Ellefson

L. M. Leeds

L. M. Leeds

L. M. Leeds (A'30-M'40) has been appointed manager, electronics laboratory of the General Electric Company's electronics department, it has been announced by Dr. W. R. G. Baker (A'19–F'26), vice-president in charge of the department. W. C. White (A'15–M'15–F'40), formerly in charge of this laboratory, has been appointed the electronics engineer of the General Electric research laboratory.

Mr. Leeds will have his headquarters in Schenectady and will have charge of electronic research and advanced development for the electronics department.

Since 1943, he has been an electronics consulting engineer for the company, as well as an expert consultant on radar and radio in the Office of the Secretary of War, with offices located in the Pentagon Building, Washington.

Mr. Leeds, born in Tulsa, Oklahoma, worked eight years in various phases of the radio industry before receiving his B.S. degree in electrical engineering from Rutgers University in 1934. He joined General Electric that year and went to work for the radio transmitter engineering department.

In 1938, Mr. Leeds was placed in charge of the development of the first General Electric television station (W2XB) in the Helderberg Mountains outside Schenectady, and the establishment of General Electric's "proving ground" station WRGB in 1939. In 1940, he was named leader of the radar-development section.

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Devereaux Martin

Devereaux Martin

Announcement was made recently by the Wilcox-Gay Corporation of the appointment of Devereaux Martin (A'36) to the engineering staff of that organization, as chief engineer.

Mr. Martin was educated at Massachusetts Institute of Technology, and has a background of fifteen years experience in the design and research division of the engineering field. He had previously been associated with Westinghouse Aircraft Division, de Forest Radio Company, Federal Telephone and Radio Manufacturing Company, Radio Receptor Company, and the J. H. Bunnell Company.

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Canadian RTPB

Reginald M. Brophy (A'25) of the Canadian Marconi Company, Montreal, has been elected president of the newly organized Canadian Radio Technical Planning Board. Ralph A. Hackbusch (A'26-M'30-F'37), vice-president of the Stromberg-Carlson Company, Toronto, was elected vice-president.

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H. S. Osborne

H. S. Osborne (A'14-M'29-SM'43) of the American Telephone and Telegraph Company was re-elected chairman of the Standards Council of the American Standards Association for 1945.

Correspondence

Correspondence on both technical and nontechnical subjects from readers of the Proceedings of the I.R.E. is invited subject to the following conditions: All rights are reserved by the Institute. Statements in letters are expressly understood to be the individual opinion of the writer, and endorsement or recognition by the I.R.E. is not implied by publication. All letters are to be submitted as typewritten, double-spaced, original copies. Any illustrations are to be submitted as inked drawings. Captions are to be supplied for all illustrations.

Determination of the Quiescent Operating Point of Amplifiers Employing Cathode Bias

The most common method by which grid-bias voltage is obtained for a receiving-type tube operating without grid current is by means of a cathode resistor, either by-passed or not depending upon the circuit requirements. A typical circuit of this type is shown in Fig. 1.

In each such application the question arises as to what the quiescent operating conditions are for a given set of circuit parameters. The usual method of solution is to draw the proper load line on the e_b-i_b characteristics for the tube, assume a value of plate current, and then revise this assumed value until a plate current is obtained which produces the value of grid-bias voltage (the drop in R_k) required to allow this plate current to flow. Such a method lacks precision and may be rather time-consuming.

A more satisfactory method than the cut-and-try process is described below. While certainly not new,[1] this method is not widely used, and a brief review may be of some assistance.

The load line in the circuit of Fig. 1 is drawn upon the basis of the equation

$$i_b = (E_{bb} - e_b)/(R_k + R_1) \qquad (1)$$

and the line obtained determines the values of plate current and plate voltage for all negative values of grid voltage. The use of such a locus is well known.

Note that in Fig. 1 the value of grid voltage e_c is dependent only upon i_b and R_k. Thus

$$e_c = -R_k i_b$$

or

$$i_b = -(e_c/R_k). \qquad (2)$$

This holds true regardless of R_1 or E_{bb}. Therefore a *grid-bias line* can be plotted on the tube characteristics by the use of (2). The desired quiescent operating point is then simply the intersection of the load line and the grid-bias line.

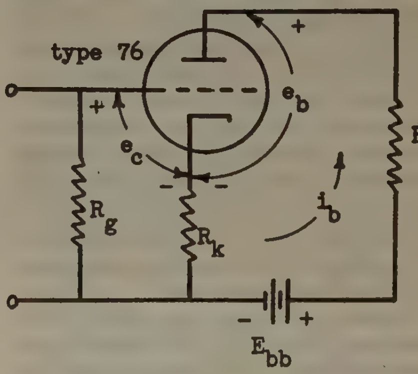

Fig. 1—Triode with cathode and plate resistors.

An example will help to illustrate the convenience of this method. Consider a type 76 triode connected as shown in Fig. 1, with $E_{bb}=250$ volts, $R_k=5000$ ohms, and $R_1=20{,}000$ ohms. The published e_b-i_b characteristics are given in Fig. 2. The usual load line, corresponding to R_k+R_1 is shown. The grid-bias line, which is, in general, *not* a straight line, is also shown, and the intersection of these two lines gives the required

[1] This method was first brought to the author's attention by W. R. Saylor of the General Radio Company, Cambridge, Mass.

quiescent operating point, at $I_{b0}=2.3$ milliamperes, and $E_{b0}=192$ volts. The grid-bias line is most conveniently drawn by assuming as values of grid bias the ones for which curves are published (such as 0, −4, −8, −12, etc., for the type 76) and calculating the plate current required by the use of (2). Thus at $e_c=-8$ volts, $i_b=8/5000=1.6$ milli-

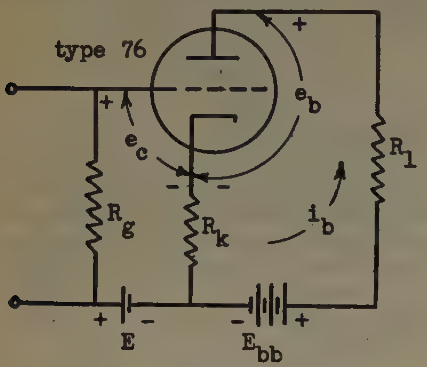

Fig. 2—Triode with additional fixed-bias voltage.

amperes, which gives the point marked X on the grid-bias line. Other points are obtained in a like manner. This is more satisfactory than to assume values of plate current and calculate the required grid bias, because interpolation is difficult between the published grid-bias values.

This method is applicable, of course, for any values of R_k and R_1. Likewise it can be applied to a circuit in which some fixed-bias voltage E either positive or negative, is introduced into the grid circuit, as shown in Fig. 3. Under such conditions (2) would become

$$\begin{aligned} e_c &= E - R_k i_b \\ \text{or} \quad i_b &= -(e_c - E)/R_k. \end{aligned} \tag{3}$$

Such an application is often useful in direct-current amplifier design.

JAMES N. THURSTON
Massachusetts Institute of Technology
Cambridge, Mass.

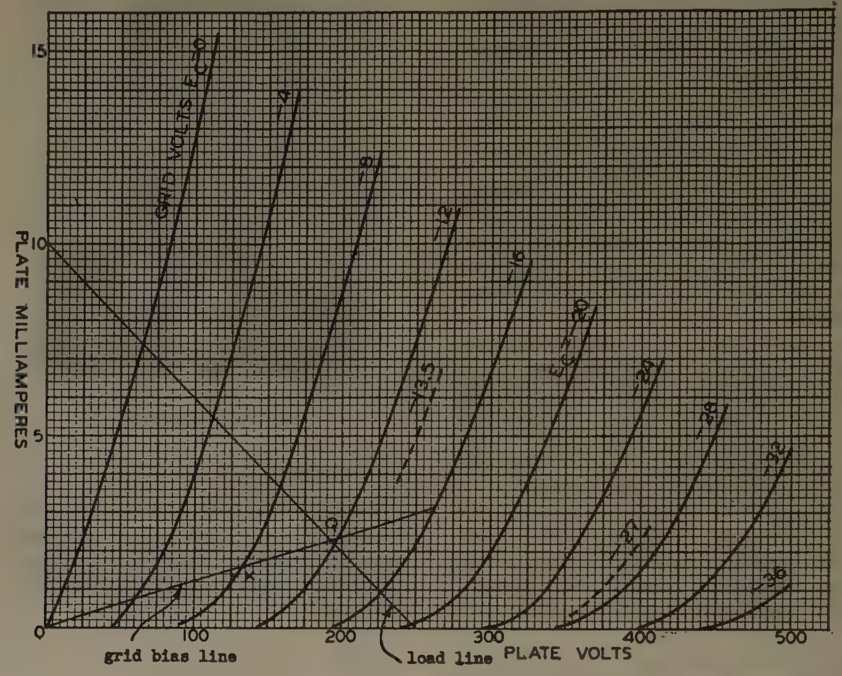

Fig. 3—Plate characteristics of type 76 triode, showing load line and grid-bias line.

Equivalent-Plate-Circuit Theorem

Mr. Preisman's letter[1] has brought out two important facts that are sometimes overlooked. These are that the equivalent plate circuit applies only to alternating voltages, currents, and associated power, and that it is strictly valid only at current amplitudes that are low enough so that nonlinear distortion is negligible. Normally the determination of power and of plate dissipation is necessary only in power or current amplifiers, in which the current amplitudes are so large that the equivalent plate circuit must be abandoned in favor of graphical methods of determining power output.

Starting from expression (6) of Mr. Stockman's letter,[2] Mr. Preisman has derived a relation, which may be stated in terms of standard symbols as follows:

$$P_i = P_p + P_0 \tag{a}$$

in which P_i is the input power to the plate circuit supplied by the direct-current source, P_p is the total plate dissipation, and P_0 is the power developed in the plate circuit (load). This relation is important not only because it may be used in determining the plate dissipation in any type of operation, but also because it shows that if the input power remains constant, which is essentially true in class A amplifiers, any reduction in power output results in increased plate dissipation. It is of interest to note that (a) follows directly from the law of conservation of energy, since it states that power supplied by the source can appear only as plate dissipation and as power developed in the load. Equation (a) can, therefore, be obtained without reference to the equivalent plate circuit or to a diagram such as Mr. Stockman's Fig. 2, which is likely to prove confusing to a student. One may obviously reverse Mr. Preisman's procedure and obtain Mr. Stockman's expression (6) from (a) under the assumptions that the direct-current resistance of the load is negligible and that the equivalent plate circuit is valid; i.e., that the plate current amplitude is sufficiently low so that nonlinear distortion may be neglected.

The question as to whether (4) of Mr. Stockman's letter should be termed a theorem or an equation would at first thought appear to be trivial. It should be noted, however, that Mr. Stockman's equation is not what is ordinarily termed the equivalent-plate-circuit theorem, but rather an equation that may be obtained from an equivalent circuit constructed by applying the general theorem to a particular type of circuit. Equation (4) does not apply to a circuit in which excitation voltage is impressed upon the plate circuit. Consider, for instance, the "reactance-tube" circuit of Fig. 1, in which the resistance r may be assumed to be so high in comparison with Z_L that Z_L is the load impedance. The plate current is not given by (4), but is

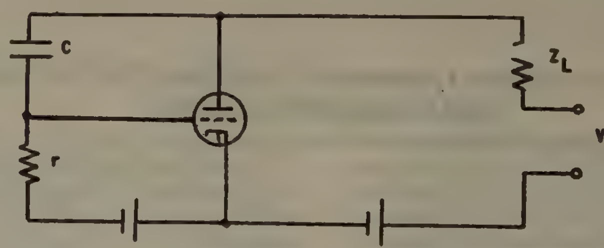

Fig. 1

[1] Albert Preisman, "Equivalent-plate-circuit theorem," PROC. I.R.E., vol. 32, pp. 642–643; October, 1944.

[2] Harry Stockman, "The validity of the equivalent plate-circuit theorem for power calculations," PROC. I.R.E., vol. 32, p. 373; June, 1944. Editor's Note: Methods of obtaining equivalent plate circuits have been discussed by Herbert J. Reich in "Theory and applications of electron tubes," second edition, McGraw-Hill Book Company, New York 18, N. Y., 1944, pp. 87–93 and pp. 671–672.

$$I_p = (V_p + \mu E_g)/(r_p + Z_L). \quad \text{(b)}$$

The equivalent-plate-circuit theorem may be stated in general form as follows: The fundamental component of alternating plate and plate-circuit currents and voltages in vacuum-tube circuits may be determined from an equivalent plate circuit in which the tube is replaced by the alternating-current plate resistance in series with a fictitious constant-voltage generator of voltage $\Sigma\mu_n E_{gn}$, or by the alternating-current plate resistance in parallel with a fictitious constant-current generator supplying a current $\Sigma g_{mn} E_{gn}$. The voltage E_{gn} is the vector sum, along any continuous path in the actual circuit, of all alternating voltages between the n'th grid and the cathode, μ_n is the mu factor of the n'th grid relative to the plate, and g_{mn} is the transconductance relating the plate and the n'th grid.

Use of the following procedure ensures the correct formation and solution of equivalent plate circuits. With suitable changes of symbols it may be applied to the circuit of any electrode.

1. (a) *Series Equivalent Circuit.* Insert the equivalent voltage μE_g and the alternating-current plate resistance r_p in series between the plate and the cathode, indicating the polarity of the equivalent voltage to be such that the cathode side of the voltage is positive, as in Fig. 2B. (The polarities indicate the vector relations between the voltages, a change in sign being equivalent to a reversal in the direction of the vector representing the voltage. The polarities may also be construed to apply to the instantaneous voltages at some instant in the cycle.) If alternating voltage is impressed upon more than one grid, there is an equivalent voltage in series with r_p for each grid that is excited.

(b) *Parallel Equivalent Circuit.* Connect the plate resistance and an equivalent constant-current generator in parallel between the plate and the cathode of the tube. The current supplied by the generator is $g_m E_g$ and should be indicated as flowing through the generator from the plate terminal to the cathode terminal. If alternating voltage is impressed upon more than one grid, there is a similar component of current for each additional grid that is excited.

2. Assume polarities for the grid and plate excitation voltages (impressed voltages) V_g and V_p. It is convenient to indicate the polarities to be such as to make the grid and the plate positive relative to the cathode. (When excitation voltages that are in phase opposition are applied simultaneously to two electrode circuits, however, obviously only one of the excitation voltages can be indicated as having such polarity as to make its electrode positive relative to the cathode.)

3. Assume positive directions for the various alternating circuit currents. It is sometimes convenient to choose the positive direction of plate current as that in which the equivalent voltage μE_g tends to cause it to flow, i.e., through r_p from plate to cathode, as shown in Fig. 2, but this is not essential.

4. Delete the tube symbol (or show dotted), all electrode supply voltages and other direct voltages, and all circuit branches not coupled to the plate (such as the screen circuit of a screen-grid tetrode, as it is usually used).

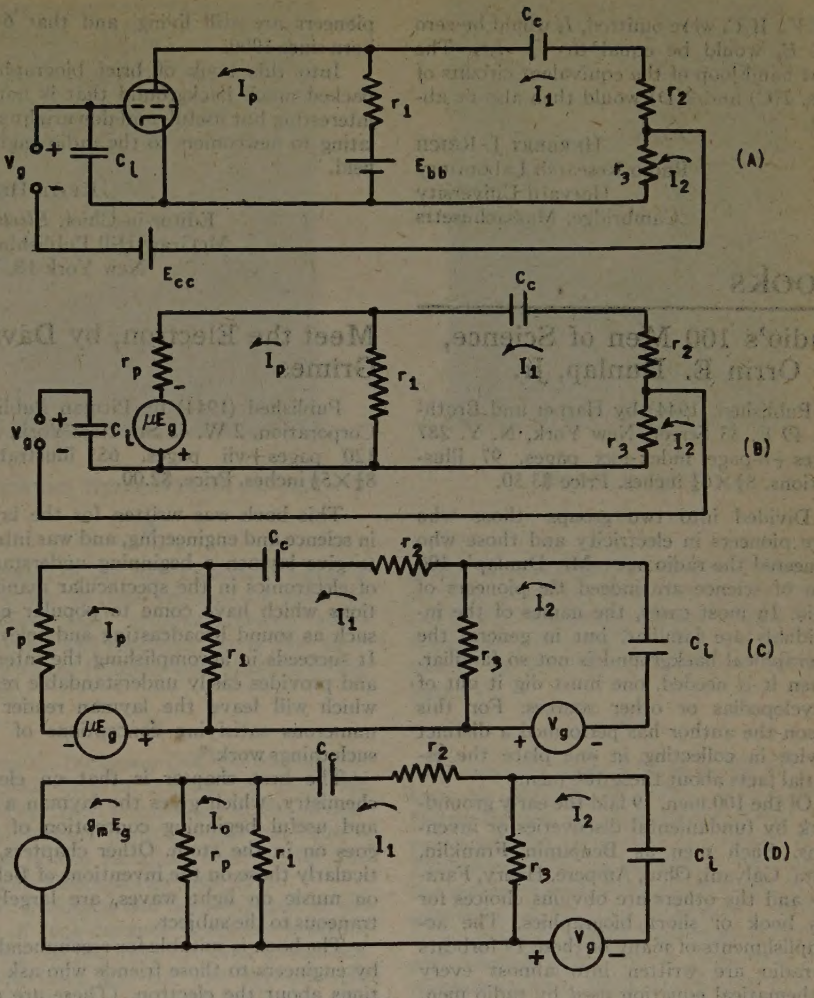

Fig. 2

5. Redraw the resulting equivalent circuit in the form in which it may be most readily analyzed.

6. Express the alternating grid voltage E_g as the vector sum of all alternating voltages between the cathode and the grid along any continuous path. This can usually be done most conveniently by reference to the original circuit. Any voltage that contributes to E_g, including the excitation voltage V_g, must be preceded by a positive sign in this summation if it tends to make the grid positive relative to the cathode, and by a negative sign if it tends to make the grid negative relative to the cathode. If alternating voltage is impressed upon more than one grid, each equivalent grid voltage in the equivalent circuit must be evaluated in this manner. It should be noted that V_g may be zero in some circuits, E_g being derived entirely from the flow of alternating currents through impedances contained in the grid circuit.

7. Write network equations for each loop of the equivalent circuit. These equations, in conjunction with the expression for E_g found in step (6) may be solved simultaneously to find all circuit currents and voltages. (Thevenin's theorem or other methods of simplifying the solution may, of course, be used.) A negative value of any current found in this manner merely indicates that the current is opposite in direction to the assumed positive value, i.e., that its phase is opposite to that assumed in the equivalent circuit.

The method of constructing an equivalent plate circuit and of evaluating E_g can be shown best with the aid of a specific example. Application of steps (1) to (4) of the above procedure to Fig. 2(A) gives the equivalent series circuit of Fig. 2(B), which may be rearranged in the more convenient form of Fig. 2(C). The parallel equivalent plate circuit is shown in Fig. 2(D). The most direct path between the cathode and the grid is through C_L. The only alternating voltage between the cathode and the grid along this path is that resulting from the flow of I_2 through C_L. Hence $E_g = -I_2/j\omega C_L$. The minus sign must be used because the flow of I_2 in the indicated positive direction causes C_L to charge in such polarity as to make the grid negative relative to the cathode. An alternative path from cathode to grid is through r_3 and V_g. Summation of voltages along this path shows that E_g is also equal to $+V_g + (I_2 - I_1)r_3$. The voltage $I_2 r_3$ is positive because the flow of I_2 in the indicated positive direction produces a voltage across r_3 that tends to make the grid positive relative to the cathode. Either of these two expressions for E_g, together with the three equations obtained by summing voltages in the three loops of the equivalent circuit, may be solved simultaneously to find the values of the currents in terms of the circuit constants

and V_g. If C_t were omitted, I_2 would be zero and E_g would be equal to $V_g - I_1 r_3$. The right-hand loop of the equivalent circuits of Figs. 2(C) and 2(D) would then also be absent.

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Books

Radio's 100 Men of Science, by Orrin E. Dunlap, Jr.

Published (1944) by Harper and Brothers, 49 E. 33 Street, New York, N. Y. 287 pages+6-page index+xx pages. 97 illustrations. 8¾×6½ inches. Price $3.50.

Divided into two groups—those who were pioneers in electricity and those who pioneered the radio age—Mr. Dunlap's 100 men of science are indeed *the* pioneers of radio. In most cases, the names of the individuals are familiar, but in general the biographical background is not so familiar. When it is needed, one must dig it out of encyclopedias or other sources. For this reason the author has performed a distinct service in collecting in one place the essential facts about these 100 men.

Of the 100 men, 19 laid the early groundwork by fundamental discoveries or inventions. Such men as Benjamin Franklin, Volta, Galvani, Ohm, Ampere, Henry, Faraday and the others are obvious choices for this book of short biographies. The accomplishments of many of these 19 forbears of radio are written into almost every mathematical equation used by radio men.

When the author chooses the remainder of the 100 names, he is in a realm in which he might find argument. After all, the selection made is but one man's opinion. It is doubtful, however, if there will be disagreement regarding his choices of those to honor. It is interesting to note that 45 of the pioneers are still living, and that 6 were born since 1900.

Into this book of brief biographies is packed much background that is not only interesting but useful, and downright stimulating to newcomers to the radio engineer's field.

KEITH HENNEY
Editor-in-Chief, *Electronics*
McGraw-Hill Publishing Co.
New York 18, N. Y.

Meet the Electron, by David Grimes

Published (1944) by Pitman Publishing Corporation, 2 W. 45 St., New York, N. Y. 120 pages+vii pages. 65 illustrations. 8½×5½ inches. Price, $2.00.

This book was written for the layman in science and engineering, and was intended to give laymen a beginning understanding of electronics in the spectacular manifestations which have come to popular notice, such as sound broadcasting and television. It succeeds in accomplishing the intention, and provides easily understandable reading which will leave the layman reader with numerous satisfying conceptions of "how such things work."

The best chapter is that on electron chemistry, which gives the layman a clear and useful beginning conception of what goes on in the atom. Other chapters, particularly those on the inventions of Bell and on music on light waves, are largely extraneous to the subject.

The book is suitable for recommendation by engineers to those friends who ask questions about the electron. (There are many such as a result of the current publicity campaign which the electron is receiving!)

The author, Dave Grimes, had the ability of clear exposition, to a degree which comes only when accompanied by sympathetic, friendly understanding of the other fellow's viewpoint. This book is a reflection of that ability which will be appreciated by his wide circle of friends in radio engineering.

ARTHUR F. VAN DYCK
Lieutenant Commander, U.S.N.R.
Washington 8, D. C.

Ultra-High-Frequency Radio Engineering, by W. L. Emery

Published (1944) by the Macmillan Company, 60 Fifth Avenue, New York, N. Y. 281 pages+13-page index+viii pages. 136 illustrations. 6×8½ inches. Price, $3.25.

This is a book on ultra-high-frequency radio engineering written for senior electrical engineering students. After an introductory chapter on the propagation of microwaves and their fields of application, there are chapters on Voltage-Regulated Power Supplies, Electronic Switching and Synchronization, Cathode-Ray Tubes and Sweep Circuits, Amplifiers, Square-Wave Testing and Transient Response, Ultra-High-Frequency Circuit Elements, Oscillators, Modulation and Detection, Radiation, and Wave Guides.

The subjects are well chosen and the information given is sound. However, the topics covered are so many and varied that it is impossible to do justice to them in a book of this size. Since this is intended as a college textbook, it is to be expected that the instructor will fill in the gaps in accordance with the needs of his students.

An interesting feature of the book is the listing of experiments to be performed by the student at the end of each chapter. Students who have the apparatus available and who perform these experiments will receive a good background in radio laboratory practice.

Any author writing a text on microwaves these days is working under a handicap because of censorship restrictions. No text on the subject written during the war can, therefore, be considered up to date.

STANFORD GOLDMAN
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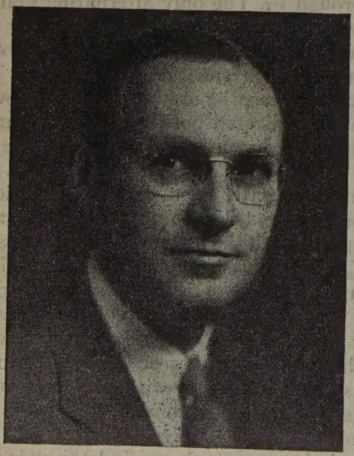
Contributors



DONALD M. BLACK

Donald M. Black (A'30-M'40-SM'43) was born on December 22, 1907, in Chatham, Ontario, Canada. He received the B.S. degree in electrical engineering in 1928 and the E.E. degree in 1933 from the University of Kansas. Since 1928 he has been a member of the technical staff of the Bell Telephone Laboratories, where he has been engaged in research and development work on radio receiving equipment in the short-wave, ultra-short-wave, and microwave region. He is a member of Tau Beta Pi, Sigma Xi, and a Member of the American Institute of Electrical Engineers.

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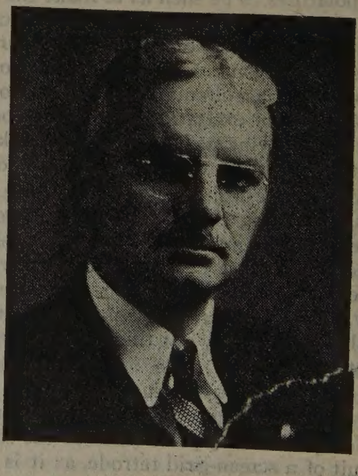
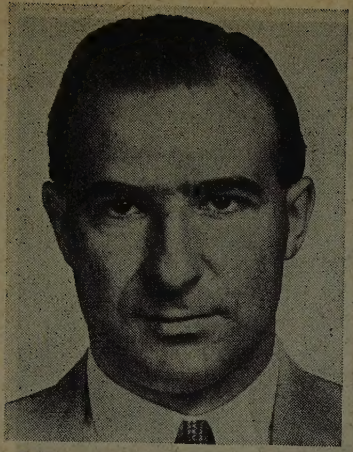
Charles R. Burrows (A'24-M'38-SM'43-F'43) was born on June 21, 1902, in Detroit, Michigan. In 1924 he received the B.S.E. degree in electrical engineering from the University of Michigan, where he had been



CHARLES R. BURROWS



ALFRED DECINO

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a research assistant. He also received the A.M. degree in physics from Columbia University in 1927; the E.E. degree from the University of Michigan in 1935; and the Ph.D. degree from Columbia University, in 1938.

Since 1924, Dr. Burrows has been a member of the radio research department of the Western Electric Company and the Bell Telephone Laboratories. His first work was in connection with the long-wave transatlantic radio telephone transmitter at Rocky Point, followed by analyses of wave propagation in the early days of short-wave radio. From 1930 to 1938 he carried out experimental and theoretical investigations on ultra-short-wave propagation. During 1939 and 1940 he was engaged in the development of radio transmitters. Dr. Burrows is a member of Sigma Xi and the American Institute of Electrical Engineers.

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Alfred Decino (A'29-M'39-SM'43) was born in Pueblo, Colorado on June 23, 1907. He received the B.S. degree in electrical engineering from the University of Colorado, 1928. From 1928 to March 1944 he was a member of the technical staff of the Radio

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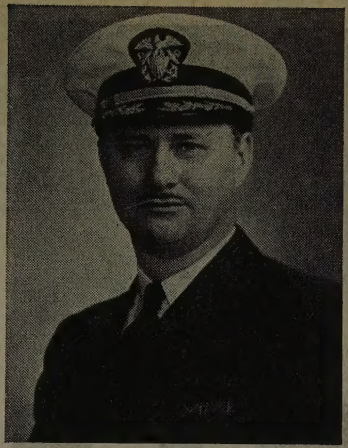
ALTON C. DICKIESON



EDWARD NELSON DINGLEY, JR.

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Department of the Bell Telephone Laboratories where he was engaged in work on short-wave and ultra-short-wave radio transmitters and receivers. He is now associated with the Hammarlund Manufacturing Company.

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Alton C. Dickieson (SM'44) was born in New York City on August 16, 1905. He studied electrical engineering at the Brooklyn Polytechnic Institute, and has been a member of the technical staff of the Bell Telephone Laboratories since 1925.

Mr. Dickieson has been engaged in the design of long-distance telephone systems, including the problems involved in connecting radio links into the wire network. His war projects have included design of wire communication and various underwater sound systems.

He is a member of the American Institute of Electrical Engineers.

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Edward Nelson Dingley, Jr., (A'28-M'43-SM'43) was born on July 18, 1902, in Kalamazoo, Michigan. He attended the Massachusetts Institute of Technology from 1922 to 1925, and received the B.S. degree in electrical engineering from George Washington University in 1936. In 1925 he was associated with RCA Laboratories as a laboratorian. He became a radio inspector for the RCA marine division in the same year, and from 1926 to 1928 was an assistant radio inspector in the Navy Yard at Washington, D.C. During 1928 and 1929, Mr. Dingley was an associate radio engineer in the Bureau of Engineering, United States Navy Department, and in 1929 and 1930 was a radio engineer with Mackay Radio and Telegraph Company.

From 1930 to 1932 Mr. Dingley was associated with the RCA Radiotron Company. He became an associate radio engineer with the Naval Research Laboratory in 1932, and transferred to the Bureau of Engineering of the United States Navy Department in 1934 where he remained until 1938. From 1938 to 1942 he was a senior radio engineer with the United States Bureau of Ships.

He was commissioned Lieutenant (junior grade) in the United States Naval Reserve in 1928, promoted to Lieutenant in 1932, and became Lieutenant Commander in 1939. After active duty in 1942 he was made Commander in 1943. He is a member of Tau Beta Pi.

ROBERT W. FRIIS

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Robert W. Friis (A'34-M'40-SM'43) was born on October 10, 1907, in Kenmare, North Dakota. He received the B.E.E. degree from the University of Minnesota in 1930. Since 1930 he has been a member of the technical staff of the radio research department of the Bell Telephone Laboratories, where he has been carrying on research and development work on short-wave and ultra-short-wave radio transmitters. Since 1941 he has been engaged in the development for production of various types of radio transmitters for the armed forces.

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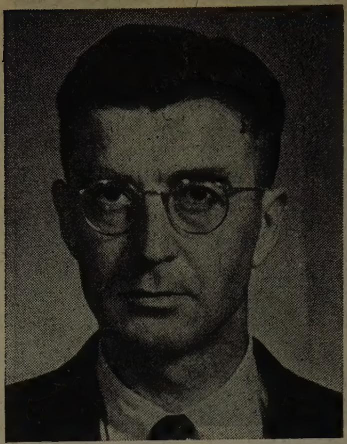
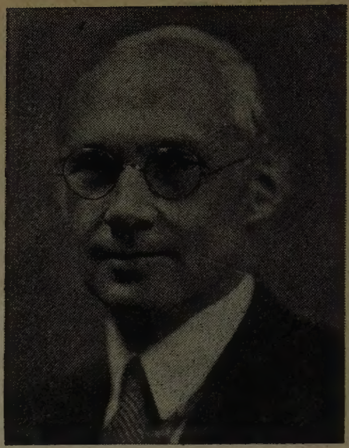
Reymond J. Kircher (A'30-M'40-SM'43) was born on November 2, 1907, in El Paso, Texas. He received the B.S. degree in electrical engineering from California Institute of Technology in 1929, and the M.S. degree in communications engineering from Stevens Institute of Technology, in 1941.

REYMOND J. KIRCHER

J. R. Pierce



M. W. Scheldorf



William C. White

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Since 1929 he has been a member of the radio research department of the Bell Telephone Laboratories engaged in development work on short-wave and ultra-short-wave transmitters and receivers. He is a member of Tau Beta Pi.

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J. R. Pierce (S'35-A'38) was born at Des Moines, Iowa, on March 27, 1910. He received the B.S. degree in 1933 and the Ph.D. degree in 1936 from the California Institute of Technology. In 1936 Dr. Pierce became a member of the Technical Staff of the Bell Telephone Laboratories, where he is engaged in electronics research.

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George Rodwin (A'25-M'40) was born in 1903 in New York City. He received the A.B. degree from Columbia University in 1923, and the E.E. degree in 1925. From 1925 to 1929 he was engaged in radio research work for the Radio Corporation of America in connection, with receiver, transmitter, and field-strength-measuring equipment. During part of 1929 and 1930 he was with the Engineering Department of the Earle Radio Corporation, developing broadcast receivers. Since 1930 Mr. Rodwin has been a member of the technical staff of the Bell Telephone Laboratories where he has been carrying on research and development work on the transatlantic radio receiving equipment and general ultra-short-wave problems.

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M. W. Scheldorf (A '26) was born on February 15, 1902, at Westside, Iowa. He received the B.S. degree in electrical engineering from Iowa State College in 1923, and then joined the Radio Department of the General Electric Company at Schenectady, N. Y. From 1930 to 1935 he was with the RCA Victor Company, at Camden, N. J. Since that time Mr. Scheldorf has been with the General Electric Company, in radio development.

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Norman F. Schlaack (A'25-M'39-SM'43) was born on June 4, 1901, at Birmingham, Michigan. He received the B.S. degree in electrical engineering from the University of Michigan in 1925. He has been a member of the technical staff of the Bell Telephone Laboratories since 1925, engaged primarily in the development of short- and ultra-short-wave transmitting equipment.

William C. White (A '15-M '25-F '42) was born in Brooklyn, N. Y., on March 24, 1890. He received the Electrical Engineering degree from Columbia University in 1912 and immediately entered the employ of the General Electric Company.

Mr. White has been identified with much pioneering work in the development of vacuum tubes. He was recently appointed electronics engineer of the research laboratory of the General Electric Company, Schenectady, N. Y.

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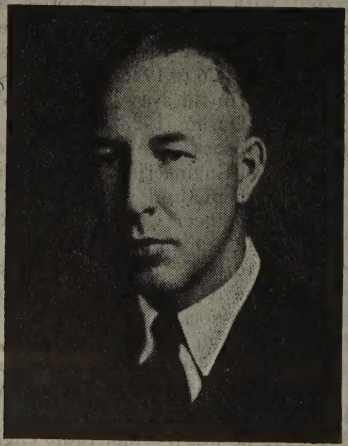
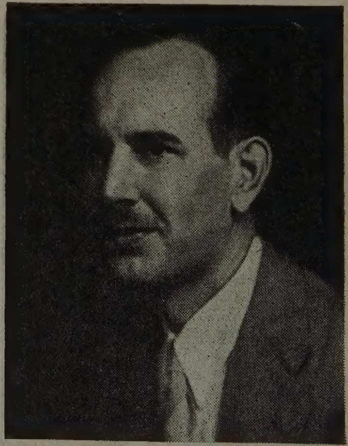
W. T. Wintringham (A'26) was born on January 18, 1904, in Brooklyn, N. Y. After receiving the B.S. degree in electric communication engineering from Harvard University in 1924, he joined the department of development and research of the American Telephone and Telegraph Company, and was closely associated with the development of the transatlantic long-wave radiotelephone system.

In 1935 Mr. Wintringham was transferred to the Bell Telephone Laboratories. His work has included investigations of the suitability of ultra-high-frequencies in crossing natural barriers, and close association in the development of the MUSA receiving system for the long-lines department of the American Telephone and Telegraph Company.

George Rodwin



Norman F. Schlaack



W. T. Wintringham